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Terms	Documents
L6 and radar$1	4

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DATE: Thursday, February 19, 2004 Printable Copy Create Case

Set Name side by side	Query	Hit Count	Set Name result set
DB=TDBD; PLUR=YES; OP=OR			
L10	l6 and radar$1	4	L10
L9	l6 and attenuat$3	13	L9
L8	l6 and l7	0	L8
L7	temperature adj compensat$3	122	L7
L6	amplifier$1 and video	188	L6
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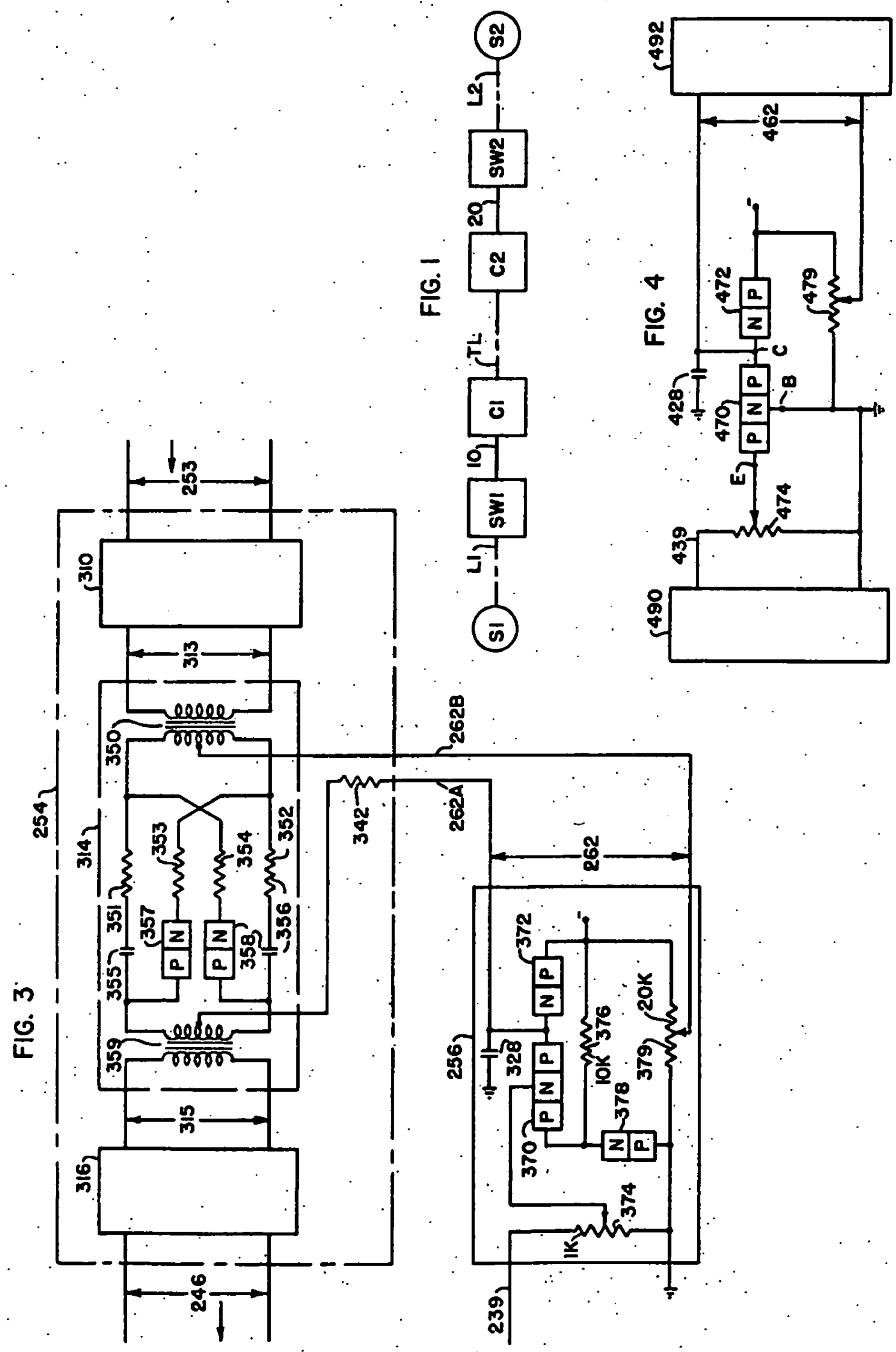
L5	l3 and radar$1	12	L5
L4	l3 and attenuat$3	23	L4
L3	l1 and l2	94	L3
L2	temperature adj compensat$3	40691	L2
L1	amplifier$1.ti,ab. and video.ti,ab.	17405	L1

END OF SEARCH HISTORY

April 8, 1958 A. R. DENZ 2,830,257

TEMPERATURE-COMPENSATED DIRECT-CURRENT TRANSISTOR AMPLIFIER

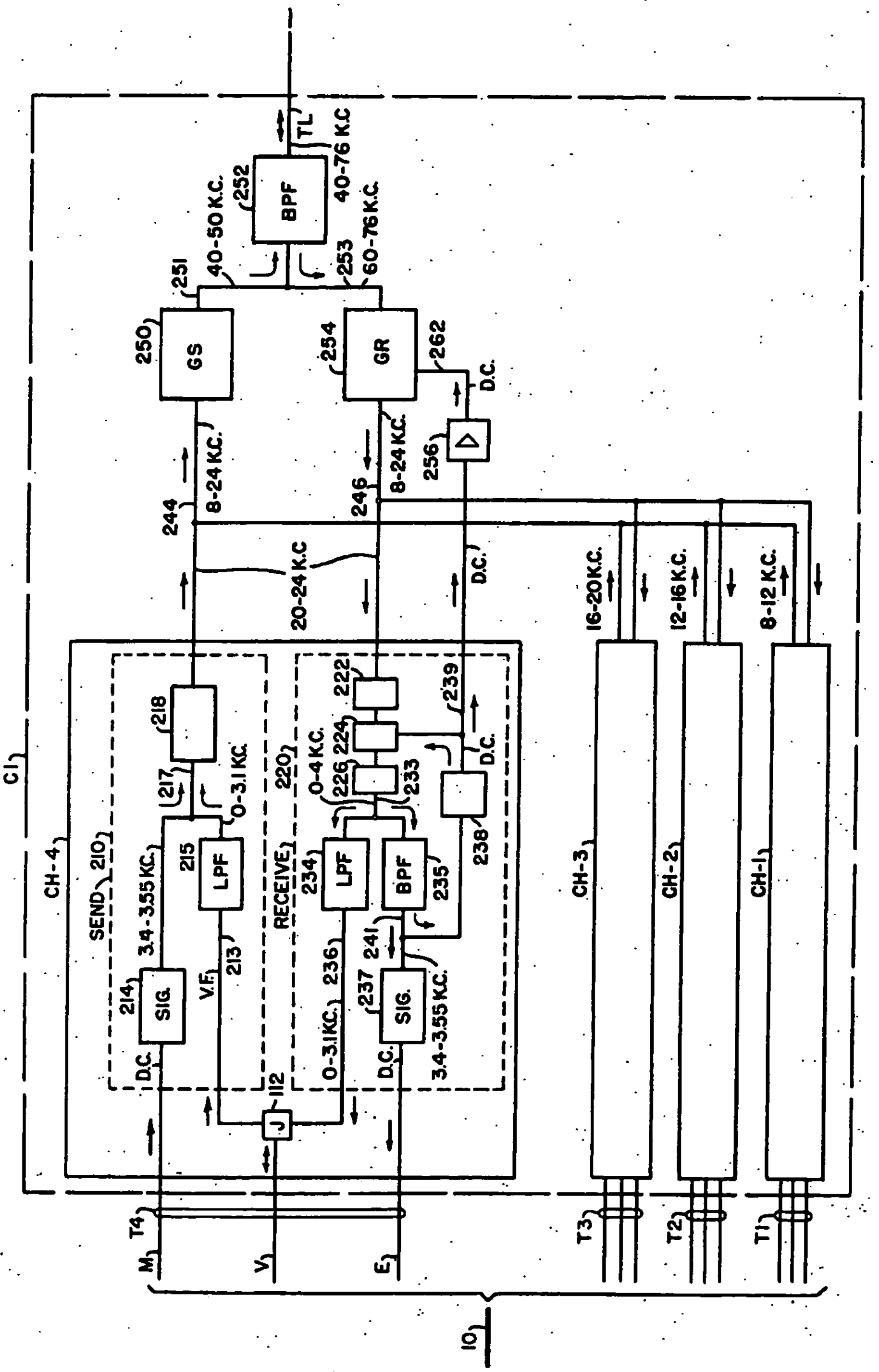
Filed July 2, 1956 2 Sheets-Sheet 1



TEMPERATURE-COMPENSATED DIRECT-CURRENT TRANSISTOR AMPLIFIER

Filed July 2, 1956 2 Sheets-Sheet 2

FIG. 2



United States Patent Office

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Patented Apr. 8, 1958

2,830,257

TEMPERATURE-COMPENSATED DIRECT-CURRENT TRANSISTOR AMPLIFIER

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Application July 2, 1956, Serial No. 595,378

6 Claims. (Cl. 323—68)

This invention relates to a temperature-compensated direct-current amplifier, and more particularly to a direct-current amplifier employing a transistor. Its object is to provide a simple and effective direct-current transistor amplifier which has zero output current at cutoff, irrespective of variations of the transistor temperature.

When a transistor is operated with the usual inverse collector bias and with current flow through the emitter diode cut off, current which increases exponentially with temperature flows through the collector diode. When current does flow through the emitter diode, the current flow through the collector diode is the sum of the cut-off current and the current which is equal to the emitter current multipled by the current-gain factor, alpha.

When a transistor is used as the current-amplifying member of a direct-current amplifier, the cutoff current is a component of the output, unless it is balanced out or compensated for in some manner. With junction transistors it is usually necessary to use a grounded emitter rather than a grounded-base configuration in a direct-current amplifier, to obtain useful current gain. Therefore, except when the emitter diode is biased to cutoff, the cutoff component of the output current is increased by the well-known regenerative characteristic of a grounded-emitter amplifier. Arrangements are known using two or more transistors, along with other circuit elements to compensate for variations of the cutoff current with temperature over the entire operating range of the amplifier. Some of these arrangements involve careful design to obtain the required result.

However, if the requirements for the amplifier are such that the cutoff current, even though regenerated by the use of a grounded-emitter configuration, may be tolerated in the output when input current is flowing; but the output must be zero when the input current is cut off, the prior arrangements are unnecessarily complicated and expensive.

According to the invention, a direct-current amplifier is provided using a single transistor, and a circuit arrangement such that the current which flows in the collector diode when the emitter diode is at cutoff, is balanced by current flow in a diode having a temperature characteristic similar to that of the collector diode.

In the preferred form of the invention, the collector diode and balancing diode are in adjacent arms of a bridge, a resistor with a sliding tap forms the other two arms, the D. C. supply source is connected in one diagonal, and the load is connected in the other diagonal.

The foregoing and other objects and features of this invention and the manner of attaining them will become more apparent and the invention itself will be best understood, by reference to the following description of an embodiment of the invention taken in conjunction with the accompanying drawings comprising Figs. 1 to 4, wherein:

Fig. 1 shows a block diagram of a telephone system including carrier equipment;

Fig. 2 shows a block diagram of a carrier terminal of Fig. 1;

Fig. 3 shows the details of a portion of the equipment of the carrier terminal including the D. C. amplifier; and

Fig. 4 shows an alternative form of the amplifier.

It has been chosen to disclose the invention as applied to a multi-channel telephone transmission system, with transmission over an open two-wire line.

Referring to Fig. 1, a switchboard SW1 at a first exchange has subscriber lines such as line L1 to station S1, and connections through conductor group 10 to carrier terminal equipment C1. The second exchange includes a similar switchboard SW2, lines such as L2 to station S2 and carrier terminal equipment C2 connected to the switchboard by conductor group 20. An open two-wire transmission line TL extends between the carrier equipments C1 and C2.

Fig. 2 shows a block diagram of the carrier terminal equipment C1. Conductor group 10 includes four trunks T1 to T4, each comprising the direct current signalling leads M and E and a voice-signal pair V. The trunks are connected to the respective channel equipments CH–1 to CH–4. As shown for channel 4, each channel equipment includes send equipment 210, receive equipment 220, and a balanced junction 112. The junction 112 includes a hybrid transformer and a balancing network for coupling the conjugate outgoing line 213 and incoming line 236 to the two-way line V. The voice signals on line 213 are coupled through low-pass filter 215 to line 217. The D. C. signals from line M modulate an oscillator in the signal equipment 214, which produces a pilot signal applied to line 217. This pilot signal is shifted between 3400 cycles and 3550 cycles, while the voice signal output from filter 215 is limited to a high of 3100 cycles. The voice and pilot signals are combined on line 217 and passed through a modulator and band-pass filter 218 to produce signals in the 20–24 kilocycle channel at the output to line 244. The outgoing signals from channel equipments CH–1 to CH–4, in separate frequency channels as indicated, are supplied in multiple to line 244 giving a four-channel band of 8 to 24 kilocycles. These signals then pass through amplifiers, modulators and band pass filters in the group-send equipment 250, where they are converted to the 40 to 56 kilocycle band, then passing over line 251 and through a group-line filter 252 to the transmission line TL.

Incoming signals on line TL, in the 60 to 76 kilocycle band, are coupled through the group line filter 252, over line 253, through the group receive equipment 254, where they are converted to signals in the 8 to 24 kilocycle band, thence to line 246 connected in multiple to the incoming lines of the channel equipments CH–1 to CH–4. The band is divided into respective channels, as indicated, by band-pass filters such as filter 222 in the channel equipment CH–4.

From the filter 222 the signals pass through an automatic gain regulator 224, and the amplifying and demodulating equipment 226, to line 233. The pilot and voice signals are then separated by the filters 234 and 235, the voice signals, which are under 3100 cycles, passing through the low-pass filter 234, while the pilot signals, which are at 3400 or 3550 cycles, pass through the band-pass filter 235. The voice signals on line 234 are then coupled through junction 112 to line V. A frequency discriminator in signalling equipment 237 converts the pilot signal to a D. C. signal on line E.

The pilot signal on line 241 is also coupled to a control amplifier and detector in equipment 238, supplying a D. C. signal on line 239 which is proportional to the amplitude of the pilot signal on line 241, to control the regulator 224. Equipment 238 also includes an

alarm circuit for indicating when the pilot signal falls below a given level. Each channel equipment includes a regulator similarly controlled. The control signal from channel 4 is further amplified by a group-control amplifier 256 for regulation in the group-receiving equipment 254.

Referring now to Fig. 3, the group receiving equipment 254 and the regulation-control amplifier 256 are shown partly in circuit detail.

Signals on line 253 from the group line filter pass through band-pass filter 310, regulator 314, and the group demodulating and amplifying equipment 316, and thence over line 246 to the receiving band-pass filters of the four channels in parallel.

The regulator 314 comprises a lattice network of resistors 351, 352, 353, and 354, blocking condensers 355 and 356, and diodes 357 and 358 in a balanced line section between transformer 350 and transformer 359, for introducing equal attenuation over the entire band according to the level of pilot signal 4. The regulator is controlled by the D. C. output on line 262 from amplifier 256 with wire 262A connected through resistor 342 to the center tap of transformer 359 and wire 262B connected to the center tap of transformer 350. The D. C. flow through the diodes 357 and 358 controls the A. C. loss of the regulator 314.

When the control signal from channel 4 is below a given level on line 239, it is desired that there be a minimum signal loss in regulator 314, and that therefore the D. C. output current from amplifier 256 to line 262 be zero. As the control signal on line 239 increases above the given level, the D. C. current on line 262 increases to increase the loss in regulator 314, thereby decreasing the signal level on line 315. Since the control signal level varies according to the level of the pilot signal 4 included in the output of the regulator 314, the system will stabilize with the current on line 262 which produces the loss in regulator 314 necessary to establish the corresponding input signal on line 239. Therefore it may be seen that the output of amplifier 256 to line 262 should increase as the input on line 239 increases above the given level, but that the exact relation between output and input is not critical.

The amplifier 256 comprises a PNP junction transistor 370, which may be type 2N44. A resistor 374 is connected from the input line 239 to ground, with a sliding tap of the resistor connected to the base terminal. A regulated one-volt negative potential is obtained at the emitter terminal by connecting it to the junction of a resistor 376 and a diode 378, which are connected in series between the negative terminal of the D. C. power supply and ground. Diode 378 may be type 6003. The control signal on line 239 has a negative potential to ground. The slider of resistor 374 is adjusted so that input current flow through the emitter diode of transistor 370 cuts off at the given signal level on line 239.

When the input signal is below the given level, the transistor is cut off by inverse bias of the emitter diode, but with the normal inverse bias potential applied to the collector terminal, cutoff current flows through the collector diode. This cutoff current increases exponentially as the temperature increases.

According to the invention, a junction diode 372, which may be a type 1N91, is connected between the collector terminal and the negative supply terminal. This balancing diode should have a variation of inverse current with temperature similar to that of the cutoff current of the collector diode of the transistor. A resistor 379 with a sliding tap is connected between the negative supply terminal and ground, forming a bridge comprising the two ends of resistor 379, diode 372, and the collector diode. The output load, comprising the D. C. path through regulator 314 in series with resistor 342, is connected between the collector terminal and the tap of resistor 379. The slider of resistor 379 is adjusted to obtain zero output current at cutoff, and remains balanced for temperature variation over a wide range.

When the input control signal exceeds the given level, current flows through the emitter diode in its forward direction, and output current flows to line 262. This output current is approximately equal to the input current at the base terminal multipled by the current-gain factor, beta. Condenser 328 filters any A. C. from the amplifier output.

Fig. 4 shows an alternative form of the amplifier using a transistor 470 with the base terminal grounded. The emitter terminal is connected to a sliding tap of a resistor 474 connected between ground and the input line 439 from a D. C. signal source 490. A junction diode 472 is connected between the collector terminal and the negative supply terminal, and a resistor 479 with a sliding tap is connected between the negative supply terminal and ground. A load 492 is connected by line 462 from the collector terminal to the tap of resistor 479. Condenser 428 filters A. C. from the output.

Operation is similar to that of amplifier 256. The transistor is cut off when the input signals on line 439 are at a negative potential. The output current is then adjusted to zero by the slider of resistor 479. When the input signals on line 439 are positive, the output current is approximately the input current at the emitter terminal multipled by the current gain factor, alpha.

While I have described above the principles of my invention in connection with specific apparatus, it is to be clearly understood that this description is made only by way of example and not as a limitation to the scope of my invention.

I claim:

1. A direct-current transistor amplifier comprising a transistor having an emitter member, a collector member, and a common base member, with emitter, base, and collector terminals connected to the respective members, the emitter and base members comprising an emitter diode, the collector and base members comprising a collector diode, a balancing diode having a variation of inverse current with temperature similar to that of the collector diode when no current is flowing through the emitter diode, supply source means for driving a direct current through the balancing diode and the collector diode in series in the inverse direction, an input circuit path through the emitter diode, an output circuit path, means including a circuit arrangement for balancing current flow in the balancing diode against current flow in collector diode for obtaining substantially zero output current at any temperature within a given range when no current is flowing through the emitter diode in its forward direction, and means responsive to input current flow through the emitter diode in its forward direction for causing corresponding current flow in the output circuit path.

2. In a direct-current transistor amplifier according to claim 1, a four-terminal bridge, with the said collector diode included in one arm, the said balancing diode in another arm, resistors in the remaining arms, the said output circuit path in the diagonal from the collector terminal to the opposite terminal, and the said supply source means in the other digonal.

3. In a direct-current transistor amplifier according to claim 1, means including a crystal diode for establishing a regulated potential of one polarity between the said emitter terminal and a reference ground, and a source of direct-current input signals of said one polarity to ground coupled to the said base terminal, whereby current flows through the emitter diode in its forward direction when the signal potential at the base terminal exceeds the regulated potential at the emitter terminal.

4. In combination with a direct-current transistor amplifier according to claim 1, a transmission path, a source of alternating-current signals including a pilot signal cou-

pled to the transmission path, a regulator included in the transmission path for reducing the variation of signal level according to the level of the pilot signal, means for obtaining a direct-current control signal which varies in value according to the level of the pilot signal at a selected point in the transmission path, means coupling the control signal to the said input circuit path, and means coupling the said output circuit path to the regulator for controlling the amount of loss to the alternating current signals passing through the regulator.

5. In a combination according to claim 4, means responsive to the said control signal having a value below a given level for blocking forward current flow through the said emitter diode, and means responsive to the control signal having a value above the given level for causing current to flow through the emitter diode in its forward direction.

6. A combination according to claim 4, wherein the said alternating-current signals pass through the said regulator before the said selected point in the transmission path.

No referencies cited.

Sept. 2, 1969 W. FISHBEIN ET AL 3,465,336

DOPPLER RADAR WITH CLUTTER CONTROLLED FILTER CHANNEL

Filed May 9, 1968 2 Sheets-Sheet 1

FIG. 1

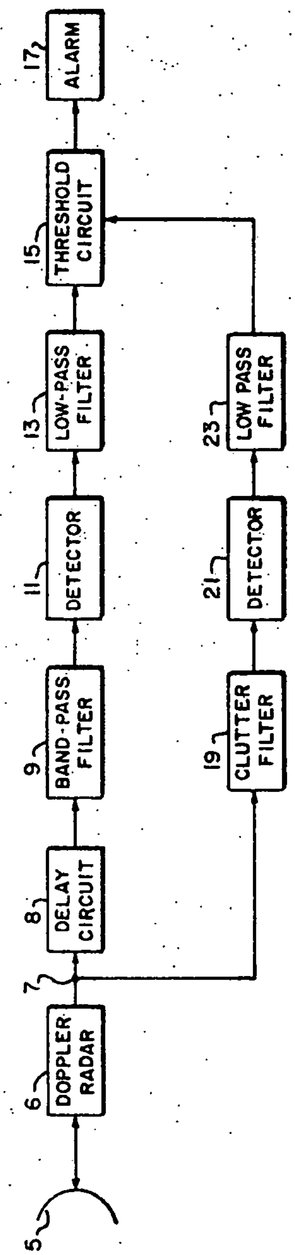
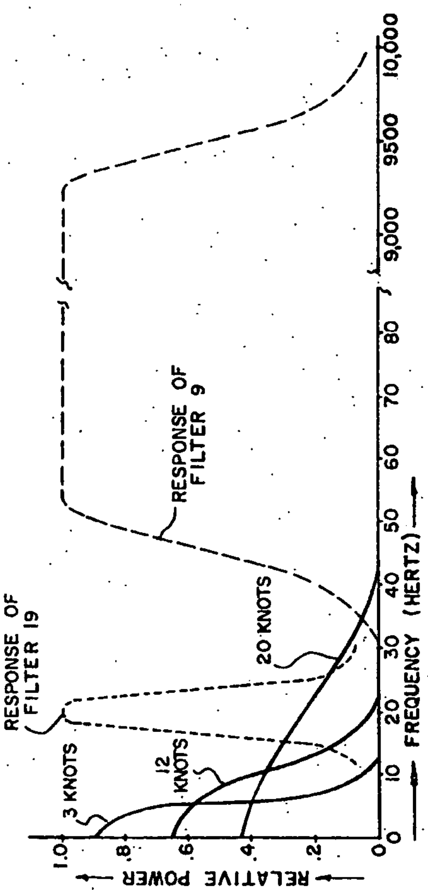


FIG. 2



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DOPPLER RADAR WITH CLUTTER CONTROLLED FILTER CHANNEL

Filed May 9, 1968 2 Sheets-Sheet 2

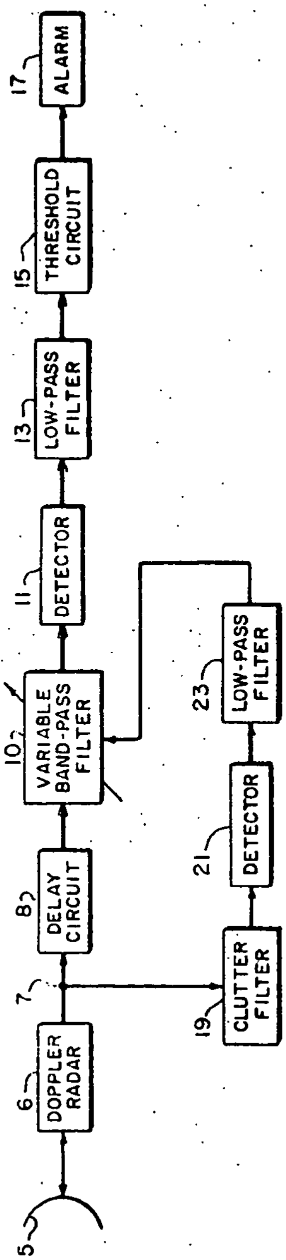

FIG. 3

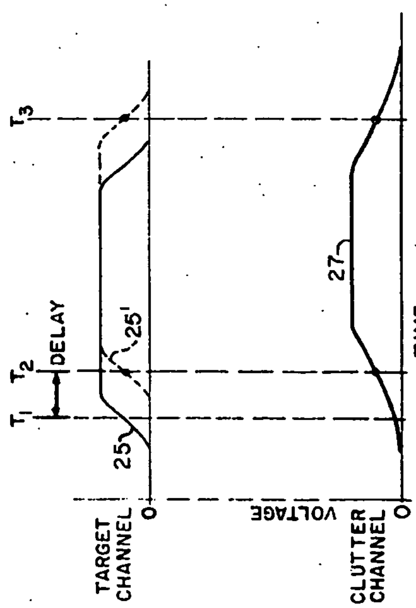

FIG. 4

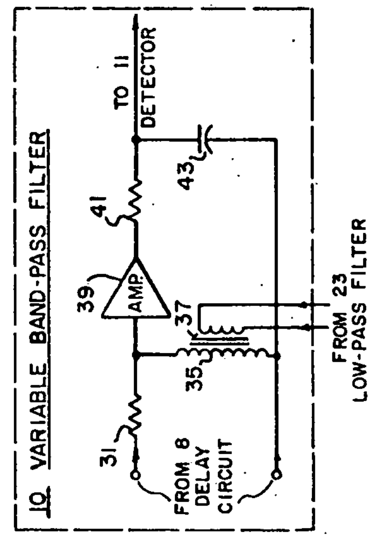

FIG. 5

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United States Patent Office

3,465,336
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3,465,336
DOPPLER RADAR WITH CLUTTER CONTROLLED FILTER CHANNEL

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Filed May 9, 1968, Ser. No. 728,016
Int. Cl. G01s 9/42
U.S. Cl. 343—7.7 2 Claims

ABSTRACT OF THE DISCLOSURE

The detected Doppler signals are applied in parallel to a target signal channel and a lower-frequency clutter control channel. The target signal channel includes an alarm which is actuated if the signal level therein exceeds the threshold of a threshold circuit. The output of the clutter control channel raises the threshold or reduces the bandwidth of the target signal channel, thus maintaining high sensitivity in the absence of wind-induced clutter and preventing false alarms caused by clutter in the target signal channel.

The invention described herein may be manufactured, used, and licensed by or for the Government for governmental purposes without the payment to us of any royalty thereon.

The present invention relates to Doppler radar sets of the type used to detect moving surface targets such as men and vehicles. When such radars are used in wooded or jungle areas, the wind-induced movement of the foliage produces Doppler clutter signals which can interfere with the detection and recognition of the desired targets. In early radars of this type, the Doppler signals were applied to earphones, which required continual monitoring by an operator. With this type of equipment skilled operators can learn to recognize different types of target and clutter signals by their distinctive sounds. Later developments eliminated the constant monitoring of the radar set by providing automatic detection systems which would actuate an alarm if a moving target appeared within the radar beam. One difficulty with such automatic systems is a reduced capability for distinguishing between desired targets and clutter compared to the continually aurally-monitored sets and the skill of the operators thereof. Wind-induced Doppler clutter signals are generally lower in frequency than the Doppler signals produced by moving men and vehicles, however high winds or brief wind gusts can produce Doppler signals of the same frequency as slowly moving men or vehicles. Filters have been inserted in the Doppler channels of such radars to block wind clutter frequencies and pass the higher frequency target signals, however if the lower frequency limit of these filters is made high enough to eliminate all wind clutter signals, some of the slower moving target signals will be lost, and if the lower limit is set low enough to include these slower moving targets, wind-induced clutter will sometimes enter the target filter and cause false alarms. The present invention comprises a Doppler radar set which overcomes these difficulties by providing an automatic Doppler detection system in which the sensitivity is automatically controlled by the amount of received clutter signal. With such an arrangement, maximum sensitivity to desired targets obtains in the absence of wind-induced clutter, and false alarms caused by wind are reduced. The circuitry includes means for applying the Doppler frequency output of the radar in parallel to a target signal channel and a clutter control channel. The target signal channel comprises a target bandpass filter followed by a detector, a low pass filter, a threshold circuit and an alarm circuit in cascade. A filter in the clutter control channel is tuned

to sense the amount of wind-induced clutter below the lower frequency limit of the target bandpass filter. The detected output of this clutter filter is utilized in one embodiment to vary the threshold or sensitivity of the threshold circuit, and in another embodiment to vary the lower frequency limit of the target filter, this achieving the desired result. A delay circuit in the target signal channel compensates for the different time constants of the two filters and further reduces the probability of false alarms.

It is thus an object of this invention to provide a Doppler radar set in which the effects of wind-induced clutter signals are minimized.

A further object of the invention is to provide a Doppler radar set in which the amount of Doppler clutter below the frequency range of desired targets is continually monitored and the main Doppler target signal channel is controlled in accordance with the amount of clutter, whereby the sensitivity of the set remains high in the absence of clutter and is reduced when clutter signals fall into the frequency range of moving targets which it is desired to detect.

These and other objects and advantages of the invention will become apparent from the following description and drawings in which:

FIGURES 1 and 3 are block diagrams of two embodiments of the invention;

FIGURE 2 is a plurality of curves illustrating the operation of the illustrative circuitry;

FIGURE 4 is a diagram showing one condition of operation of the illustrated circuitry; and

FIGURE 5 is a detailed circuit diagram of the bandpass filter 10 of FIGURE 3.

The Doppler radar set 6 and antenna 5 of FIGURE 1 comprise circuitry for transmitting and receiving either a pulsed or continuous wave radar beam, detecting any moving targets within said beam by known techniques to produce Doppler frequency signals indicative thereof at output lead 7. The Doppler signals are applied in parallel to a target signal channel and a clutter control channel, the former comprising delay circuit 8, bandpass filter 9, detector 11, low pass filter 13, threshold circuit 15 and alarm circuit 17, all in cascade. The clutter control channel comprises clutter filter 19, detector 21, and low pass filter 23, the output of which controls the threshold or sensitivity of circuit 15. FIGURE 2 shows in dashed lines the frequency responses of the two filters 9 and 19, as well as the amount and frequency range of wind-induced clutter for sustained winds of 3, 12 and 20 knots. It can be seen that the filter 9 has a lower half power point of about 45 Hertz and an upper half power point of about 9,500 Hertz, this being the target frequency range of interest. The lower frequency limit of 45 Hertz corresponds to a target radial velocity of approximately 2.5 kilometers/hr., for X band radar with a wavelength of approximately 3 cm. It can be seen that the Doppler clutter signals produced by winds of 12 knots or less all fall below the frequency range of the target frequency filter 9, however winds of 20 knots or more will produce Doppler clutter within the target signal frequency range. The clutter bandpass filter 19 has its center frequency at approximately 20 Hertz and a bandwidth such that clutter frequencies caused by winds which are too low to overlap the response of target filter 9 will cause negligible response in the clutter control channel. Thus, while winds of 12 knots will produce some response in the clutter control channel, the amplitude of this response is not sufficient to produce any appreciable control function. The passband of filter 19 and the gain of the clutter control channel is arranged so that any wind clutter signals of such frequency range as to overlap the target signal channel, such as the 20 knot curve illustrated, will produce the desired control

function at the output of the clutter control channel. It should be noted that the foliage velocity on the average is always much lower than the wind that causes it. This accounts for the relatively low Doppler clutter frequencies caused by winds of substantial force. Returning to FIGURE 1, the target signal channel without the clutter control channel comprises a prior art type of automatic detection Doppler radar referred to above. In such a system the target Doppler frequencies of interest are selected by filter 9, rectified or detected by detector 11, converted to a smoot DC voltage by low pass filter 15 and applied to the control input of threshold circuit 15, which in the prior art had a fixed or manually adjustable threshold or sensitivity. When the output of filter 13 exceeded the threshold of circuit 15, the alarm 17 would be actuated to notify personnel in the area that the radar had picked up a moving target. In accordance with the present invention, the threshold or sensitivity of the circuit 15 is made automatically responsive to the output of the clutter control channel, so that in the presence of large amounts of wind-induced clutter, the threshold of circuit 15 is raised to de-sensitize this circuit for the duration of the clutter. Thus, during periods of little or no wind, the threshold of 15 may be such that the system is extremely sensitive to the desired targets, but during the presence of wind, the clutter which may leak into the target signal channel will not produce false alarms, since the same clutter is utilized to raise the threshold level. The detector 21 and low pass filter 23 function in the same way as the corresponding circuitry of the target signal channel. The threshold circuit 15 may comprise merely a back-biased diode, the amount of back bias determining the threshold thereof. The output of the clutter control channel would provide the variable back-bias. Such threshold circuits are shown in the Gillmer Patent 3,140,486, issued on July 7, 1964. The rest of the circuitry within the blocks of FIGURE 1 is conventional and need not be described in detail.

The purpose of the delay circuit 8 is to compensate for different transit times of the same signal through the two channels, which can cause false alarms. Since the clutter filter 19 has a lower center frequency and a smaller bandwidth than target filter 9, it will require a longer time for the energy to build up and decay in this filter than in the target channel filter 9. Thus the filter 19 acts as a delay cicuit and the delay circuit 8 is inserted to equalize the delay in both channels. This effect of a differential delay can be illustrated by the diagram of FIGURE 4. The pulses 25 and 27 represent the outputs of the two low pass filters 13 and 23 respectively in the absence of the delay circuit 8, for a wind clutter input signal of such frequency that it enters both channels. It can be seen that this clutter signal 25 in the target channel will reach the threshold circuit 15 before the same signal in the clutter control channel has had a chance to raise the threshold. Thus the leading edge of the pulse 25 will trigger a false alarm. The delay circuit 8 is given a delay of t_2 minus t_1, equal to the differential delay of the two channels in the absence of delay circuit 8, so that the same clutter signal in both channels will reach the threshold circuit simultaneously. The dashed-line curve 25' is the delayed clutter signal of the target channel. The time delay of t_2 minus t_1 is determined from the time difference required for the two curves 25 and 27 to reach one half of the maximum voltage of each.

In the second embodiment of FIGURE 3 a threshold circuit with a fixed or manually variable threshold is utilized, as in the prior art, but the output of the clutter control channel is utilized to automatically raise the lower frequency limit of a variable bandpass filter 10 of the target signal channel in response to the presence of output from the clutter signal channel. Thus in the absence of wind-induced clutter, the low frequency cutoff of filter 10 would be the same as that of filter 9, as illustrated in FIGURE 2. Winds of 20 knots or more would generate sufficient output from the clutter control channel to automatically raise the lower frequency limit of filter 10 far enough so that the amount of wind clutter signal in the target signal channel would be reduced by such an amount that the clutter signal would not be of sufficient amplitude to overcome the threshold of circuit 15. Thus the result is the same as in the circuit of FIGURE 1, the output of the clutter channel automatically de-sensitizing the target signal channel, but accomplishing this by reducing the bandwidth thereof. The circuit elements of FIGURE 3 bearing the same reference numerals as those of FIGURE 1 perform the same function. FIGURE 5 shows in detail how the bandwidth of the filter 10 may be controlled in the desired manner. The variable bandpass filter may comprise an RL network comprising a series resistor 31 and a shunt inductor 35, which network comprises a high pass filter with a cutoff frequency which varies directly with the magnitude of the inductance of 35. Inductor 35 is part of a saturable reactor, the inductance of which is controlled by the amount of control current fed to the control winding 37 thereof from the output of low pass filter 23, which is the output of the clutter control channel. The inductor 35 and resistor 31 are proportioned so that in the absence of current in the control winding 37, the frequency response of the filter would be the same as that of the filter 9 shown in FIGURE 2. Clutter signals will cause the inductance of 35 to decrease, thus raising the cutoff frequency (or the lower frequency limit) of the filter sufficiently so that the clutter signal is kept out of the target signal channel. The voltage across inductor 35 is applied to an amplifier 39 and thence to a low pass filter comprising a series resistor 41 and a shunt capacitor 43. The combination of the low pass filter 41, 43 and the high pass filter 31, 35 comprises the variable bandpass filter. The values of 41 and 43 are chosen to produce the fixed upper frequency limit of approximately 9,500 Hertz, as shown in FIGURE 2. The amplifier 39 in addition to producing gain, decouples the high and low pass filters from each other. The fixed filter 9 of FIGURE 1 may be similar to the filter of FIGURE 5, but with fixed elements. It should be noted that saturable reactor comprising windings 35 and 37 are shown in schematic form only. In practice the windings 33 and 37 would be applied to a two-window core with the control winding 37 on the center leg of the core, and the winding 35 on the two outer legs. Instead of a variable reactor, the lower frequency limit of the filter may be controlled by varying the resistance of 31. This can be accomplished by utilizing a resistor 31 of semiconductor material, which has a high temperature coefficient of resistance, and applying the control signal from low pass filter 23 to a heating element adjacent to the resistor 31. The change in resistance then produces the required change in the response of the filter.

While the invention has been described in connection with illustrative embodiments, the inventive concepts disclosed herein are of general application, hence the invention should be limited only by the scope of the appended claims.

What is claimed is:

1. A Doppler radar set having means to actuate an alarm when moving targets enter the beam of said set, comprising, means to derive Doppler frequency signals from moving targets and moving clutter within said beam, a target signal channel and a clutter control channel, means, to apply said Doppler signals in parallel to both said channels, said target signal channel being tuned to a band of frequencies representing the Doppler shifts of desired targets and said clutter control channel being tuned to a band of frequencies below that of said target signal channel, said means to actuate an alarm being connected in said target signal channel, and means to de-sensitize said target signal channel in response to output from said clutter control channel, and wherein said target signal channel includes a threshold circuit with a control in-

put for automatically changing the threshold thereof and wherein said means to de-sensitize said target signal channel comprises a connection from the output of said clutter control channel to the control input of said threshold circuit.

2. A Doppler radar set having means to actuate an alarm when moving targets enter the beam of said set, comprising, means to derive Doppler frequency signals from moving targets and moving clutter within said beam, a target signal channel and a clutter control channel, means to apply said Doppler signals in parallel to both said channels, said target signal channel being tuned to a band of frequencies representing the Doppler shifts of desired targets and said clutter control channel being tuned to a band of frequencies below that of said target signal channel, said means to actuate an alarm being connected in said target signal channel, and means to de-sensitize said target signal channel in response to output from said cutter control channel, and wherein said target signal channel comprises a variable bandpass filter including a control input for automatically changing the lower frequency limit of the bandpass thereof, said variable bandpass filter normally being tuned to said band of frequencies, and wherein the output of said clutter control channel is connected to said control input.

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HERBERT C. WAMSLEY, Assistant Examiner

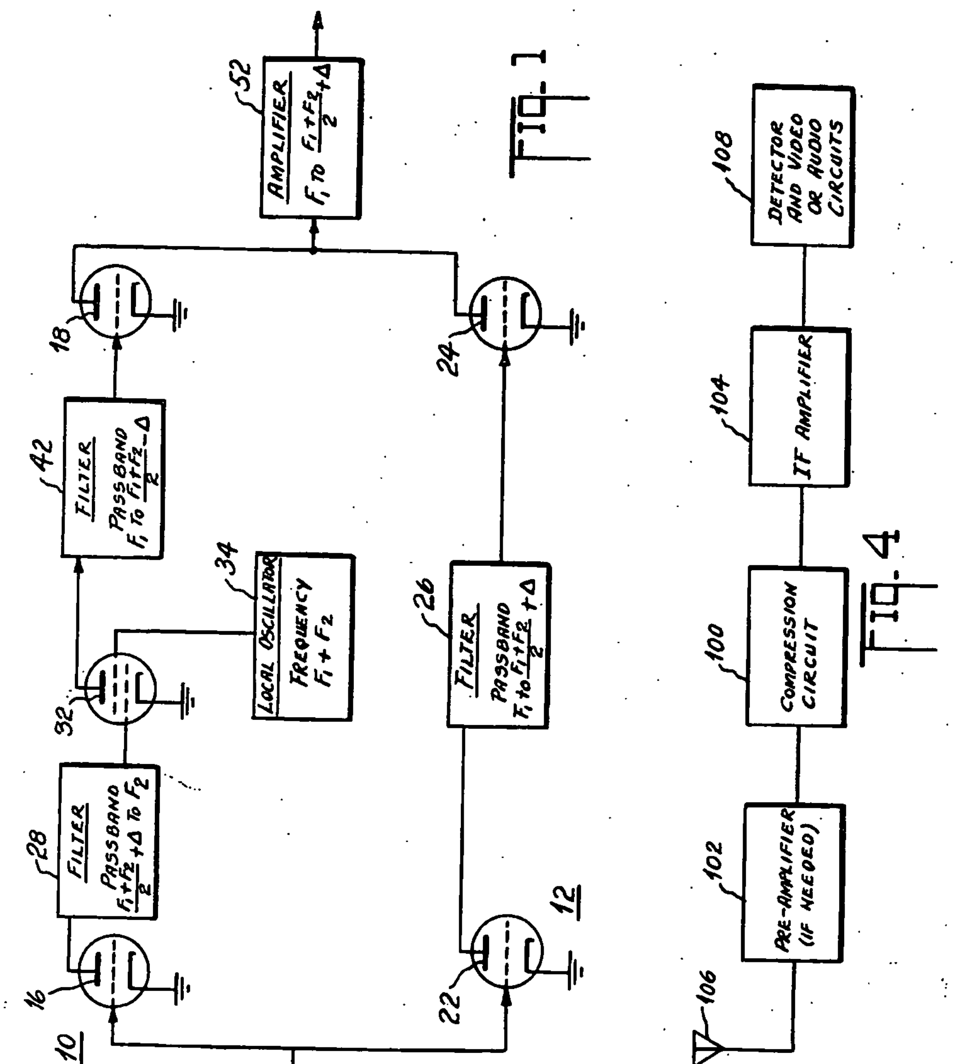
U.S. Cl. X.R.

340—258

June 7, 1960 N. FREEDMAN ET AL 2,939,918

SYSTEM FOR COMPRESSING BANDWIDTH

Filed Aug. 24, 1954 2 Sheets-Sheet 1



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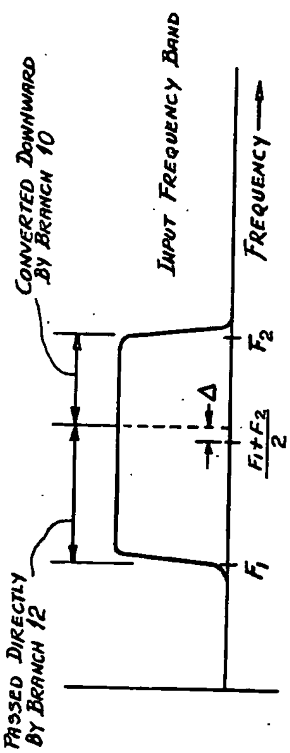

FIG. 2

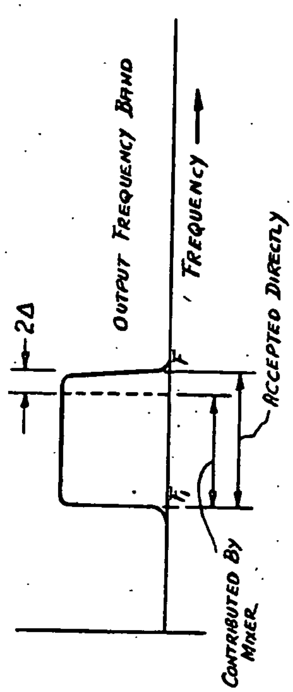

FIG. 3

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2,939,918

SYSTEM FOR COMPRESSING BANDWIDTH

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7 Claims. (Cl. 179—15.55)

This invention concerns a method and a system for compressing bandwidth, and more particularly, it concerns a method and a system for compressing the intermediate frequency bandwidth necessary in wide-coverage receivers whose intermediate frequency bandwith is wider than is necessary for fidelity reasons alone.

This invention is particularly adapted for use in a radar beacon receiver. The wide bandwidth of a radar beacon receiver is made necessary by the fact that interrogating transmitters may scatter in frequency over a considerable region. This invention is adapted for use with any receiver which is required to listen in simultaneously on a wide band of frequencies for possible transmissions, although the signals themselves only occupy a narrow spectrum bandwidth. There is no inherent frequency limitation in this system; however, its specific advantages are of most use in high-frequency, very wide-band receivers.

An object of this invention is to provide a method for compressing bandwidth.

A further object is to provide a method for compressing intermediate frequency bandwidth in a wide-coverage receiver to slightly more than one-half its original value.

A further object is to provide a system for compressing bandwidth.

A further object is to provide a system for compressing intermediate frequency bandwidth in a wide-coverage receiver to slightly more than one-half its original value.

Other objects and many of the attendant advantages of this invention will be readily appreciated as the same becomes better understood by reference to the following detailed description when considered in connection with the accompanying drawings wherein:

Fig. 1 is a schematic wiring diagram of a preferred embodiment of this invention with most components shown in block form,

Fig. 2 shows a graphical plot of the bandwidth input to the circuit of Fig. 1 and its division by the branches of Fig. 1,

Fig. 3 is a graphical plot of the output frequency band of the circuit of Fig. 1, and

Fig. 4 is a block diagram of a receiver including this invention.

The embodiment of the invention shown in Fig. 1 includes a pair of isolated circuit branches 10 and 12. The circuit branch 10 includes a decoupling vacuum tube 16 at its input end and a decoupling vacuum tube 18 at its output end. The circuit branch 12 includes a decoupling vacuum tube 22 at its input end and a decoupling vacuum tube 24 at its output end. The decoupling vacuum tubes of the circuit branches 10 and 12 serve to decouple the circuit branches from each other to prevent interaction between the components in the circuit branches. The input ends of circuit branches 10 and 12 are adapted to be connected to a signal source 25; signal source 25 provides a narrow spectrum signal anywhere in a selected, very wide, acceptance frequency band extending from frequency F_1 to frequency F_2 (Fig. 2). The circuit branch 12 includes a filter 26. The filter 26 is characterized by a passband which includes the lower half of the bandwidth accepted from signal source 25 plus a small increment (Δ) of the uper half. The passband of filter 26 extends from frequency F_1 to

$$\frac{F_1+F_2}{2}+\Delta$$

The filter 26 is connected directly between the output end of decoupling vacuum tube 22 and the input end of decoupling vacuum tube 24.

The circuit branch 10 includes a filter 28. The filter 28 is characterized by a passband in the acceptance band of the circuit which permits passage of the frequencies from signal source 25 not passed by filter 26. The passband of filter 28 is equal to

$$\frac{F_1+F_2}{2}+\Delta$$

to frequency F_2. Circuit branch 10 further includes a mixer stage 32. A local oscillator 34 is connected to one of the grids of the mixer stage 32. The output end of filter 28 is connected to another of the grids of mixer stage 32. The frequency generated by the local oscillator 34 is equal to the sum of the frequencies defining the limits of the acceptance band of signals derived from signal source 25. The mixer stage 32 heterodynes the signals from filter 28 and from local oscillator 34. A third filter 42 is connected in the plate circuit of the mixer stage 32. The filter 42 is characterized by a passband which accepts the difference beat frequency from mixer stage 32. The bandpass of filter 42 extends from frequency F_1 to

$$\frac{F_1+F_2}{2}-\Delta$$

Filter 28 operates with a sharp low frequency cutoff. Filter 42 operates with a sharp high frequency cutoff. The output end of filter 42 is connected to the input end of decoupling vacuum tube 18. Decoupling by tubes is desirable at least at one end so that there can be no undesirable effects due to feedback around the loop comprising the two circuit branches 10 and 12.

An amplifier 52 is connected to the output ends of circuit branches 10 and 12. The passband of amplifier 52 is only slightly more than one-half the acceptance bandwidth of the branches 10 and 12 to signals from source 25. The outputs from circuit branches 10 and 12 are combined and amplified by amplifier 52. The output of the amplifier is an amplified signal, which may be detected, just as in a conventional system. If desired, more complex filters may be substituted to afford the isolation above afforded by the isolation vacuum tubes 16, 18, 22, and 24, making the latter unnecessary. In other words, more involved band-dividing filters may be designed which split up an acceptance band in two sections while maintaining a constant input impedance. These are similar to the cross-over networks used in high-and-low frequency loudspeaker installations. The use of these might eliminate the necessity for the input pair of vacuum tubes. There is shown in Fig. 4 in block form, the relationship of this compression circuit to the remainder of a receiver. The compression circuit 100 in accordance with this invention is connected between a pre-amplifier 102, if needed, and an IF amplifier 104. The pre-amplifier 102 is connected to antenna 106. The IF amplifier is connected to detector and video or audio circuits 108.

The operation of the circuit is best illustrated by a numerical example. An inspection of the quantities involved in this numerical example shows that the mixing action in the circuit branch 10 converts the upper half

of the circuit acceptance band to signals from source 25 into the lower half of the band from source 25. To illustrate, let it be assumed that the acceptance frequency band extends from 10 to 90 megacycles. Further, let it be assumed that Δ is 5 megacycles. Therefore, the passband of filter 26 extends from 10 to 55 megacycles. The passband of filter 28 extends from 55 to 90 megacycles. The local oscillator frequency is equal to 10+90 or 100 megacycles. The difference generated in the mixer stage 32 is equal to 100−(55 to 90) or 10 to 45 megacycles. As a result, the 10 to 90 acceptance frequency band from source 25 is fitted into a band of 10 to 55 megacycles. If the frequency increment Δ were not employed in the filter design, the acceptance band of signals from source 25 would be divided exactly in half. The lower half of the band would go through circuit branch 12 and the upper half of the band would go into the mixer stage 32 in circuit branch 10. Under this condition of operation a frequency at or close to the center of the input band would arrive at the output of circuit branches 10 and 12 since filter characteristics cannot be made infinitely sharp. The output of circuit branches 10 and 12 would then contain frequency components which are very close together. Frequency components which would be close together would beat in the output detector to produce undesirable low frequency modulation on the desired signal. Therefore, by staggering the branch acceptance bands by the increment Δ the acceptance band center area is passed only by branch 12, thereby precluding undesirable low frequency modulation on the desired signal due to the beat effect in the output detector. The magnitude of Δ selected depends upon the excellence of the filters; it also depends upon the harmful effects of a low frequency beat in using the receiver.

The spurious low-frequency modulation may be reduced to whatever extent necessary by making Δ large enough. How large it needs to be depends on the width of the signal spectrum, and the sharpness of the filter characteristics.

The output frequency response will be flat only if the gain through both circuit branches 10 and 12 are equal. This may conveniently be arranged by setting the gain of a tube in the circuit. If the gain through all filters is unity, it follows that the mixer conversion gain is unity for flat frequency response.

The bandwidth is compressed to make possible amplification over a wide band using a reasonable number of tubes. Gain-bandwidth considerations would force use of a very large number of tubes in a conventional IF design of 80 megacycles bandwidth, for example. With half the bandwidth, the number of tubes is cut by much more than half.

The system described may be cascaded for further bandwidth narrowing. However, too much cascading generally lowers receiver sensitivity.

Obviously many modifications and variations of the present invention are possible in the light of the above teachings. It is therefore to be understood that within the scope of the appended claims the invention may be practiced otherwise than as specifically described.

We claim:

1. A bandwidth compression circuit for a predetermined frequency band comprising; a first circuit branch, a second circuit branch, means for connecting said circuit branches in parallel and for isolating said circuit branches against interaction with one another; said first circuit branch including a first filter characterized by a passband corresponding to the lower part of the predetermined frequency band, the passband differing from one-half the predetermined frequency band by an amount delta; said second circuit branch including a second filter characterized by a passband equal to the difference between the predetermined frequency band and the passband of said first filter, both of said filters manifesting sharp cutoff at the separation frequency, a local oscil-

lator for generating a frequency which is equal to the sum of the frequencies defining the limits of the predetermined frequency band, a mixer stage connected to said second filter and to said local oscillator, and a third filter connected to the output of said mixer stage and characterized by a passband for passing only lower sideband frequencies from said mixer stage; the amount delta being no larger than is necessary for preventing harmful low frequency modulation of the desired signal.

2. A bandwidth compression circuit for a predetermined frequency acceptance band comprising a signal source of said band, an amplifier, parallel circuits connecting said source to said amplifier, each of said circuits including therein two decoupling stages one near the beginning and one near the end thereof, one of said circuits between its said stages having a filter passing only signals of frequencies in the range of a lower part of at least half of said frequency band plus an increment equal to a small fraction of the entire frequency range of said acceptance band, and the other of said circuits having sequentially in series therein, beginning immediately after its said first decoupling stage, a filter passing only signal frequencies in the range from said lower part of said band plus said increment to the upper limit of said band, then a mixer tube, and then a filter passing only signal frequencies in the range of said lower part minus said increment, and an oscillator connected to said mixer tube and heterodyning the signal frequencies passing between the filters in said other circuit to a range within the said lower part of said acceptance band.

3. The circuit as set forth in claim 2 wherein said lower part is substantially half of the entire acceptance frequency band.

4. A bandwidth compression circuit for a predetermined frequency acceptance band comprising a signal source of said band, an amplifier, parallel circuits connecting said source to said amplifier, one of said circuits having a filter passing only signals of frequencies in the range of the lowest frequency in the band to at least half of said frequency acceptance band plus an increment equal to a small fraction of the entire frequency range of said acceptance band, and the other of said circuits having sequentially in series therein first a filter passing only signal frequencies in the range from said half of said acceptance band plus said increment to the upper limit of said band, and then a filter passing only signal frequencies in the range from the lowest frequency in said acceptance band to said about half of said acceptance band minus said increment, and means connected to said other of said parallel circuits between said filters in that parallel circuit, for heterodyning the signal frequencies passing the first filter in that said other parallel circuit with a frequency equal to the sum of about the lowest and highest limits of said acceptance band to a range within the said lower part of said acceptance band.

5. A bandwidth compression circuit for a predetermined frequency acceptance band comprising a signal source of said band, an amplifier, parallel circuits connecting said source to said amplifier, one of said circuits having a filter passing only signals of frequencies in the range of a lower part of at least half of said frequency acceptance band plus an increment equal to a small fraction of the entire frequency range of said acceptance band, and the other of said circuits having sequentially in series therein first a filter passing only signal frequencies in the range from said lower part of said band plus said increment to the upper limit of said band and then a filter passing only signal frequencies in the range of said lower part minus said increment, and an oscillator connected to said other of said parallel circuits for heterodyning the signal frequencies passing the first filter in that said other parallel circuit with a frequency equal to the sum of the limit frequencies of said acceptance band.

6. The circuit as set forth in claim 4, and means associated with the beginning and end of each of said parallel circuits for decoupling them.

7. The circuit as set forth in claim 5, and means associated with each of said parallel circuits for decoupling them.

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United States Patent
McKinney

[15] 3,701,154

[45] Oct. 24, 1972

[54] **MATCHED FILTER**

[72] Inventor: **Edward J. McKinney**, Arlington, Va.

[73] Assignee: **The United States of America as represented by the Secretary of the Navy**

[22] Filed: **March 9, 1971**

[21] Appl. No.: **122,403**

[52] U.S. Cl. ....**343/18 E**, 340/3 R, 343/17.2 R, 343/17.2 PC

[51] Int. Cl. **G01s 7/36**, G01s 9/23

[58] Field of Search.........343/17.2 R, 17.2 PC, 18 E; 340/3 M

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Primary Examiner—T. H. Tubbesing
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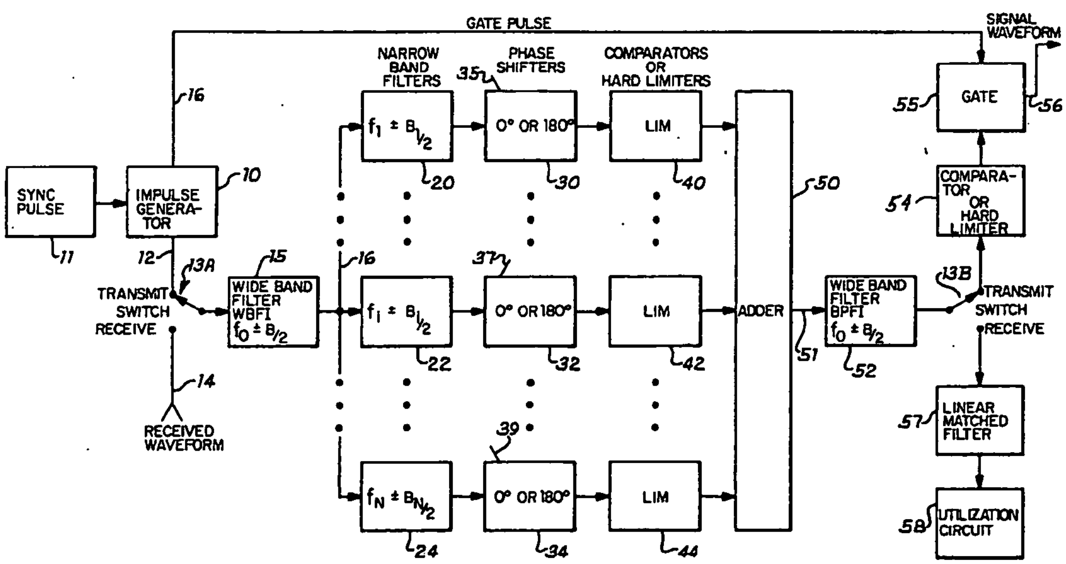
[57] **ABSTRACT**

A matched filter having a frequency band signal input which is coupled through a wide band filter, through a plurality of narrow band filters coextensive in frequency with the wide band filter to phase code and decode the signal, through a similar number of hard limiters or comparators to an adder, and through a band pass filter, all between transmit-receive (T-R) switches of an electromagnetic echo ranging target detection system to detect targets in noise interference and countermeasure environments.

3 Claims, 2 Drawing Figures



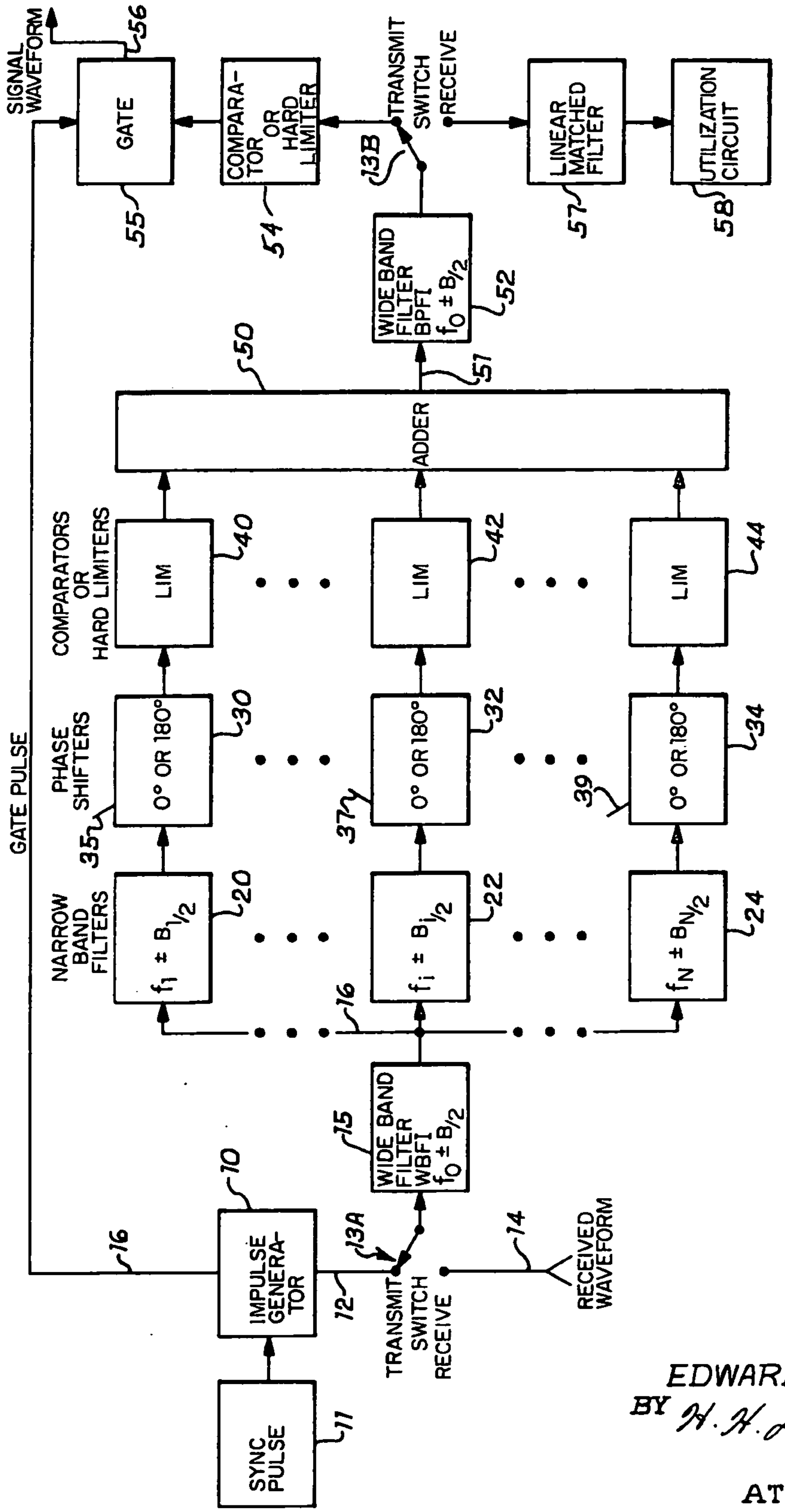


Fig. 1

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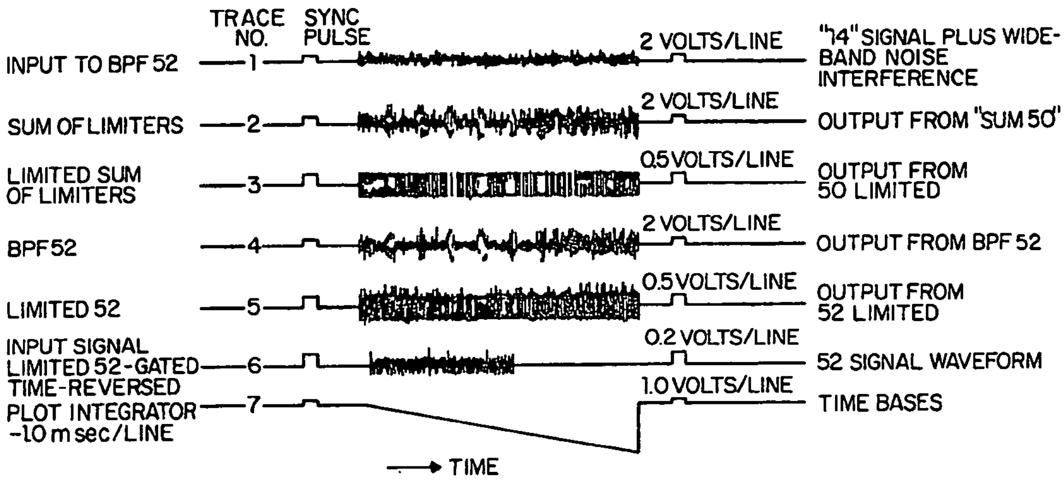

SIMULATION WAVEFORMS

Fig. 2

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MATCHED FILTER

BACKGROUND OF THE INVENTION

This inventi n relates to matched filter techniques and m re particularly to matched filters used in the transmitter and receiver circuits of an echo ranging system with phase shifting means to phase code and decode the transmitted and received frequency signals, respectively, and with limiting means to control the false alarm rate to detect real targets in a low signal-to-noise ratio environment and where countermeasures are received back to provide false ranges of the target.

In the prior art, in order to perform its basic task, an echo ranging system must detect the presence of a signal echo in the presence of background interference and properly identify the detected signal. Signal detection requires the system to recognize a small increase in the power received in the presence of a high level of background interference due to natural phenomena and to jamming type countermeasures (CM). In an automatic (unmanned) system, this requires pre-establishment of a threshold or set of thresholds which, when exceeded, indicate the presence of a signal. Due to the statistical nature of the interference, there will always be a finite probability that the interference alone will exceed the threshold requirements giving rise to a false alarm. The false alarm probability and the probability of detection of the signal are both dependent on the threshold settings and are thus, for a given system, completely interdependent. In general, threshold settings are determined by the maximum tolerable false alarm probability and the resulting probability of detection accepted as the best attainable performance for the given system.

In the presence of countermeasures, the background interference can have a wide variety of waveforms as well as a wide range of amplitude levels. A system with fixed thresholds will be unable to maintain the false alarm probability at the tolerable level unless some means is provided for normalizing the input interference to the level for which the thresholds were set. AGC circuitry has been proven inadequate for this purpose in the countermeasures environment where the interference may have a wide variety of spectral characteristics over a period of time.

A desirable system characteristic would be the maintenance of a constant false alarm rate (at the minimum tolerable level) for all interference waveforms at all amplitude levels. In general, this is not attainable. An acceptable system characteristic is the maintenance of a constant false alarm rate for all amplitude levels of a given interference waveform and a limited variation in false alarm rate for a wide variety of interference waveforms. When the false alarm rate for a wide variety of interference waveforms (at all anticipated amplitude levels) can be maintained between pre-established limits, the system will be said to exhibit controlled false alarm rate (CFAR).

Given a maximum false alarm probability which will be assumed to be controllable, the problem then becomes one of establishing and evaluating the best receiver configuration in the countermeasures environment. The best receiver is assumed to be the one which provides the highest probability of detection for a given signal-to-interference ratio. An alternate definition of the best receiver is that receiver which requires the

lowest signal-to-interference ratio for a 50 percent probability of detection. If the receiver has CFAR capability, the two definitions are equivalent.

The simplest model, which has been extensively used in detection studies, considers an environment in which the target echo is received together with stationary additive Gaussian noise. Under these conditions, conventional normalization techniques, such as AGC, are effective in providing CFAR and the optimum detector is then a matched filter or one of its equivalents, such as an active correlator. When the interference is non-Gaussian and/or nonstationary, these techniques become vulnerable to increased false alarms. Interference wavefroms which are rapidly swept in frequency and narrow band (compared to the signal bandwidth) interference waveforms are particularly effective in causing increased false alarm rates in such systems. A simple limiting counter-countermeasure (CCM) technique has been developed which solved the normalization problem in the presence of swept and wide band interference waveforms. This technique does not provide any enhancement of the signal but is compatible with conventional matched filtering. The technique was susceptible to highly colored interference waveforms and, in particular, to CW interference waveforms. Depending on the relative frequencies and bandwidths of the signal and interference, this susceptibility is in the form of lowered detection probability and increased false alarm probability or a decrease in both the detection and false alarm probabilities.

A patent issued to G. M. Kirkpatrick, U.S. Pat. No. 3,112,452, discloses a CCM technique which provides CFAR capability in the presence of a wider variety of interference waveforms and, in particular, against high colored interference waveforms such as CW or pulses of continuous sine waves. The Kirkpatrick technique does not provide any signal enchancement but is compatible with pulse compression systems, and, in fact, has been successfully applies to a system employing a linear frequency modulated pulse (CHIRP). The present invention is an extension of the techniques disclosed in this patent which provides further signal enhancement without loss of CFAR capability and with some increase in CFAR performance possible by means of an adaptive feature of the invention.

SUMMARY OF THE INVENTION

In the present invention a CCM matched filter circuit is placed in the transmit receive (T-R) switch circuit of an echo ranging system, such as a radar or sonar system, so that the transmitted signals and the received signals pass through the same filter circuit. The input to the matched filter has a wide band filter to allow a predetermined bank of frequencies through and these frequencies are split into several channels by a plurality of narrow band filters. The output of each of the narrow band filters is passed through a phase shifter to phase code and decode the narrow bands of frequencies and then hard limited to establish thresholds of the signal to unify signal amplitudes. All of the channeled frequencies are summed or added and the output of the adder applied through a band pass filter of the same band as the wide band filter to the T-R switch. Transmitted signals are passed through a comparator or hard

limiter and a gate to the transmitter antenna while the received signals are passed through a single linear matched filter to the point of use, such as a display device. The phase shifters in the matched filter circuit provide the means to phase code the transmitted signals which phase decode th received or ech signals in like manner. Accordingly, it is a general object of this invention to provide a CCM matched filter circuit capable of splitting a wide band frequency signal into a plurality of narrow band frequency signals, phase code and decode these narrow band signals, hard limit these signals, and sum the hard limited and coded signals to normalize the transmitted and received echo signals for higher resolution target detection without loss of CFAR capabilities.

BRIEF DESCRIPTION OF THE DRAWING

These and other objects, advantages, and features will become more apparent to those skilled in the art as a more detailed description proceeds when considered along with the accompanying drawing in which:

FIG. 1 is a block circuit schematic of the CCM matched filter of this invention; and

FIG. 2 provides traces of the waveforms appearing at various points in FIG. 1.

DESCRIPTION OF THE PREFERRED EMBODIMENT

Referring more particularly to FIG. 1 an impulse generator **10** triggered by sync pulses from a source **11** generates short pulses on the output **12**, the time duration of which is short compared to the impulse response of a wide band filter (WBFI) to which these pulses are applied through a transmit-receive switch (T-R) switch **13A**. The receiver terminal **14** of the T-R switch **13A** is coupled to the output of a sensor (not shown) such as a sonar transducer or radar antenna. The output of the wide band filter **15** is approximately its impulse response and is a small central portion of the Sin X/X envelope which is fed simultaneously to a bank of narrow band filters **20–22–24**, represented herein by frequencies $f_1 \pm B_{1/2}$, $f_i \pm B_{i/2}$, $f_N \pm B_{N/2}$ which represent a bank of narrow band filters of any desirable number between **20** and **22** and between **22** and **24**. The total frequency span of these narrow band filters should span the band B of the wide band filter **15** although they may not necessarily cover the band. That is, the narrow band filters **20–22–24** may be contiguous or have frequency gaps between adjacent filters. Noncontiguous narrow band filters are desirable in a sonar application to reduce the effects of reverberation. In any application the CCM matched filter theoretical performance improves as the number of narrow band filters is increased. In practice, a compromise will be required between the number of narrow band filters and the frequency gaps between filters based upon the maximum practical band width B of the input wide band filter **15** and the minimum practical band width of the narrow band filters.

The outputs of the narrow band filters **20–22–24** are fed individually to a bank of two-state phase shifters **30–32–34** which provide either 0° phase shift or 180° of phase shift selected by the switch means **35, 37,** and **39.** Thus the phase shifters either pass the narrow band filter outputs unaltered (0° phase shift) or inverts their

polarity by 180° phase shift. The purpose of thes phase shifters **30–32–34** is to provide the syst m designer with a means of phase coding to adjust the system performance to th environment of the sonar or the radar system as most desirable.

The outputs f the bank of phase shifters **30–32–34** are fed individually to a bank of comparators or hard limiters **40–42–44**. A comparator is a two state device whose output voltage is either plus or minus some value of voltage V depending on the polarity of the input waveform, as well understood by those skilled in the art. In theory the performance of a comparator and an ideal hard limiter are identical. Accordingly, the limiters **40–42–44** each limit the narrow band frequency in either its zero or 180° phase condition in their channels **20,30,40; 22,32,42;** and **24,34,44** in their respective channel outputs.

The outputs of the comparator-limiter **40–42–44** are then summed in an adder circuit **50** of any well known construction and the sum produced on the output **51**. Since the output of each limiter or comparator **40–42–44** is either plus or minus V, the output **51** voltage from the adder **50** is restricted to the range plus or minus NV where N is the number of narrow band channels. The output **51** of the adder **50** is passed through a band pass filter **52**, the band width B of which is the same as the band width of the wide band filter **15**. The output of the band pass filter **52** is applied to the second section **13B** of the T-R switch and in its transmit switch position is passed through a comparator or hard limiter **54**, thence through gate **55** to an output **56** which is adapted to be coupled to the transmitter antenna or transducer. The gate circuit **55** is gated by a gate pulse from the impulse generator **10** by way of a conductor means **16**. The output of band pass filter **52** is hard limited in **54** to produce a waveform which varies in time between two voltage stated V_1 depending on the polarity of the waveform out of band pass filter **52**. This two state waveform is time gated by the gate circuit **55** in accordance with the triggers applied thereto over the conductors **16** from input pulse generator **10**. If the frequency separation between all adjacent channels is the same and the gate time duration set as an integral multiple of this frequency separation, the gated waveform is used directly as the transmitted waveform. Otherwise, the gated waveform, which may be digitally represented, is time reversed and the result used as the transmitted waveform.

The T-R switch **13A,13B** switches alternately in unison after each transmitted pulse from the transmit position to the receive position electronically, as is well understood in the operation of sonar and radar ranging devices. In the receive positions of the T-R switch **13A,13B** interference alone or signal plus interference is applied over the input conductor **14** to the T-R switch **13A** which proceeds through the wide band filter **15** and on to the system of narrow band frequency channels in the same manner as impulses from the generator **10** in the transmit position. In the receive position of the T-R switch **13B** the output of the band pass filter **52** is conducted through a linear matched filter **57** to a utilization circuit **58** illustrated herein by block which may be of any type such as a cathode ray tube indicator. Since the received signal is, by virtue of symmetry or reversal, a time reverse of the impulse

response of the system, the system is matched to the signal waveform in the sense that its response to the signal exceeds its response to any other wavef rm. It is not a matched filter in a conventional sense of the word due to the nonlinearities n both transmit and receive functi ns. Since the narrow band filter channels do not overlap, the voltages V out of the channel limiters **40–42–44** are independent. Therefore, with interference only at the input to the system, the limiter voltages add noncoherently and the root mean square (RMS) noise power out of the adder **50** is proportional to NV^2 where N is the number of channels and V is the limiter output voltage level. If the input to the system consists of signal alone, the voltage output of the limiters **40–42–44** will add coherently and produce a peak power proportional to $(NV)^2$. Thus, the maximum attainable signal-to-interference ratio will be proportional to the number of channels N.

OPERATION

In the operation of the CCM matched filter illustrated in block in FIG. **1** with reference to the traces of FIG. **2**, let it first be assumed that the T-R switches **13A,13B** are in the transmit position as shown in FIG. **1**. In this switched position a sync pulse will trigger the impulse generator **10** to start a time base running as illustrated in trace number **7** of FIG. **2** (not illustrative of any circuit results). The impulse generator **10** will produce a pulse of frequencies having an envelope of the center of Sin X/X which will be conducted through the wide band filter **15** and the several channels to the adder circuit **50**. The output of the adder **50** would be the sum of the frequencies of the narrow band channel filters **20–22–24** plus higher order harmonics generated by the hard limiters **40–42–44**. The output of band pass filter **52** would be hard limited, time reversed and transmitted by way of gate **55** with the waveform as shown in trace number **6**. This waveform was obtained by recording the gate output and then reversing the tape developed therefrom to provide a transmit waveform. This transmitted impulse of frequency would be coded in accordance with the setting of the switches **35–37–39** in the phase shifters **30–32–34** in accordance with the desired phase coding for the environment in which the sonar or radar system is operating. This trace, trace number **6**, with wide band noise added to simulate interference, echoed back and received via **14** produces the signal plus wide band noise interference as shown in trace number **1**.

After transmission the T-R switches **13A,13B** will be switched to the receive mode, as well understood by those skilled in the sonar and radar range art, to receive an echo signal of the transmitted signal as shown by trace **1** with wide band noise added to simulate interference. The received signal will be conducted through the wide band filter **15** and through the several filter channels to the adder circuit **50**. Since the transmitted signal was coded, this reflected coded echo signal will be decoded by the phase shifts **30–32–34**. The output of the adder circuit **50**, as shown by trace **2**, will be conducted through the wide band filter **52**, as shown by trace **4**, and through the linear matched filter **57** to the utilizati n circuit such as a cathode ray tube display to indicate whether the target object is real and not a target object caused by countermeasure. Additional traces in FIG. **2** which were made during simulation tests f r information purposes which d not represent waveforms at any point in the circuit of FIG. **1** are trace number **3** which is the limited output of adder **50** in the receive mode and trace number **5** which is the limited output of wide band filter **52**. The receive signals will be pulse compressed. Since interference signal powers will add noncoherently as NV^2 and target signal powers will add coherently by $(NV)^2$ a target signal will be very pronounced while interference or countermeasure signals will be suppressed in the signal-to-interference ratio proportional to the number N. The phase coding and decoding of the transmitted and received signals operate to provide the best CFAR performance to the environment anticipated. For this CCM matched filter there is no necessity to maintain a linear phase function over the entire frequency band *B* of the filter bank although it may be desirable to maintain linear phase over each individual channel. The functioning of the CCM matched filter herein described provides excellent signal-to-interference noise ration in adapting its interference performance to the anticipated environment and the subsequent definition and generation of a transmitted signal based on the anticipated environment by the selected coding of the phase shifter **30–32–34** to the 0° or 180° phase conditions by switches **35–37–39**. In this way the environment establishes the parameters of the above-described CCM matched filter.

While many modifications may be made in the constructional details without departing from the preferred embodiment shown herein to acquire similar results and functions, I desire to be limited in the spirit of my invention only by the scope of the appended claims.

I claim:

1. A counter-countermeasure matched filter circuit comprising:
   - an input of frequency signals from a T-R switch coupled for alternate switching to a transmitter pulse generator and to a receiver output;
   - a wide band filter in said input to pass frequency signals in a predetermined band of frequencies to an output thereof;
   - a plurality of narrow band filters coupled in common to the wide band filter output and each having an output;
   - a two state phase shifter coupled to the output of each narrow band filter for selectively shifting the phase of each narrow band of frequency 0° and 180° on an output thereof providing phase coding and decoding of said wide band of frequencies;
   - a limiter coupled to the output of each narrow band filter to limit the amplitude of the narrow band frequency on an output thereof;
   - an adder network having inputs coupled to the outputs of said limiters to sum the narrow band frequency signals into a wide band of frequencies coextensive in bandwidth to said wide band filter on an output thereof, each circuited narrow band filter, phase shifter, hard limiter, and adder network providing narrow band filter channels; and
   - a band pass filter coupled to th utput of said adder circuit to filter out harmonics produced by the addition of said narrow bands of frequencies with an output coupled through said T-R switch to switch

for transmission and reception, said switch for transmission being through a hard limiter and gate and said switch for reception being through a linear matched filter to a point of use whereby transmitted frequency signals are phase coded in said plurality of narrow band filters in a plurality of narrow frequency bands and the received echo signals are phase decoded in said plurality of narrow band filters to eliminate false alarms and countermeasure signals not conforming to the frequency coded signals.

2. A counter-countermeasure matched filter circuit as set forth in claim 1 wherein

said T-R switch in one switched condition couples said pulse generator output through said filter channels, through said hard limiter, and through said gate for phase coded transmission of pulses, and in the other switched condition couples the receiver through said filter channels f r phase decoding the echo pulses and through the linear matched filter for target echo use.

3. A counter-countermeasure matched filter circuit as set forth in claim 2 wherein

said plurality of narrow band filters have separate adjacent narrow frequency bands within the frequency limits of said wide band filter, said narrow frequency bands ranging from contiguous to noncontiguous gaps in frequency therebetween to control the effects of reverberation.

* * * * *

United States Patent [19]
Teich

[11] **3,875,399**
[45] **Apr. 1, 1975**

[54] **MULTI-FREQUENCY OPTIMUM HETERODYNE SYSTEM**

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[21] Appl. No.: **407,603**

Related U.S. Application Data

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[52] **U.S. Cl.** **250/199**, 325/17, 325/47, 328/133, 343/7.7, 336/28

[51] **Int. Cl.** **H04b 9/00**

[58] **Field of Search** .... 325/15, 17, 44, 47; 328/133, 141; 343/8, 9, 7 A, 7.7, 13 R, 12 R, 12 A; 250/199; 356/28

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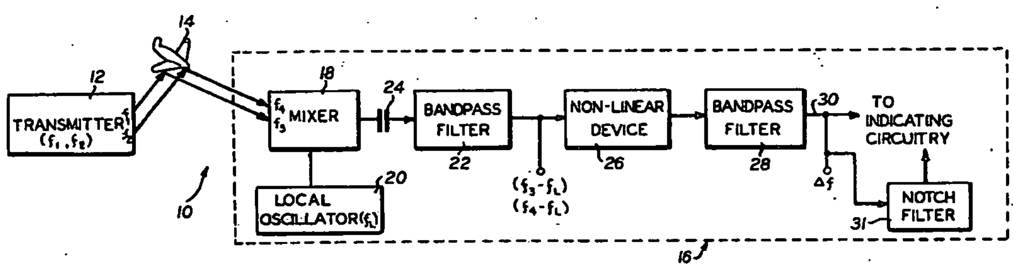
[57] **ABSTRACT**

A transmission system which transmits first and second frequency signals separated by a difference frequency to a moving target which reflects or scatters the Doppler-shifted signals as respective third and fourth signals which are separated in frequency substantially by the difference signal. The third and fourth signals are heterodyned in a receiver with a local oscillator, and the output of the heterodyne stage is connected to a filter which passes preselected ones of the heterodyned signals to a non-linear device. The non-linear device produces a signal in response to the preselected signals, the frequency of which is substantially equal to the difference frequency. The transmission may be directly to a receiver for communicating so that the Doppler shift occurs primarily because of relative movement between the transmitter and receiver. The receiver may also be used separately for detection of Doppler-shifted multiple signals radiated by remote or local excited species in order to detect and identify the excited species.

22 Claims, 2 Drawing Figures



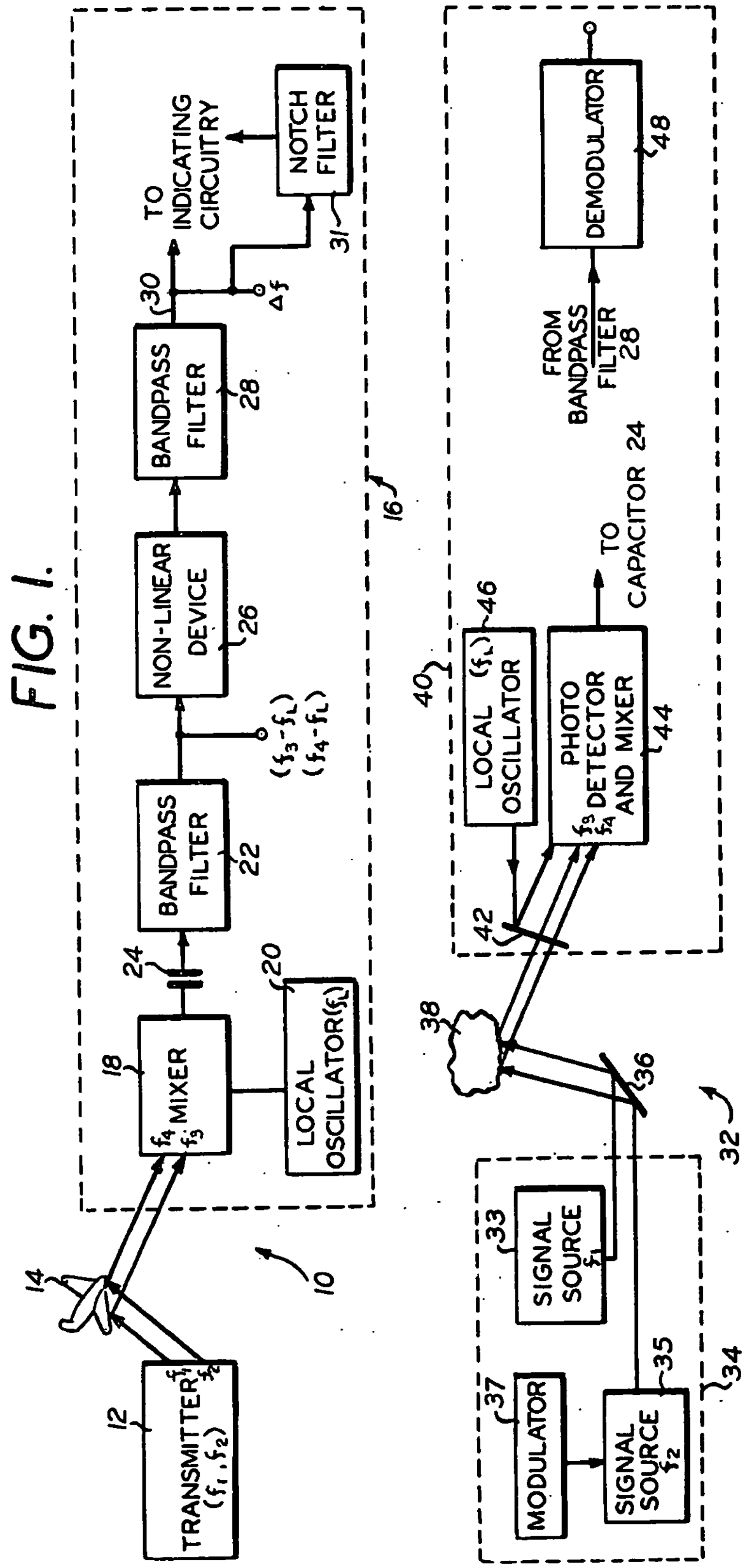


FIG. 1.

FIG. 2.

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MULTI-FREQUENCY OPTIMUM HETERODYNE SYSTEM

This is a continuation of application Ser. No. 069718, filed Sept. 4, 1972.

This invention relates generally to a system which is ideally suited for use as an acquisition or a tracking radar or communication system, and to detect and identify a source of remotely or locally emitted weak signals.

It is a well known fact in radar technology that when a target to be detected or monitored is moving with respect to the radar receiver, the signal reflected or scattered from the target will be shifted in frequency due to the Doppler effect. Moreover, this so-called Doppler shift is related to velocity. Accordingly, when the velocity of the target is known the Doppler frequency shift can be easily calculated. Accordingly, the receiver may then be turned to the frequency shifted signal to increase the sensitivity of the system. However, a major problem is presented when the target velocity is unknown which, in turn, places a severe limitation on presently available acquisition and tracing radar systems.

To be more specific, when the velocity of the target is unknown, as is the case with most target detecting systems, the Doppler shift in frequency of the reflected signal cannot be calculated. In order to compensate for the unknown frequency shift, the receiver circuits are usually made broadband so that they can accommodate signals having frequencies which fall into a large band. However, increasing the bandwidth of the receiver circuits also increases the amount of noise signals which can pass through the receiver to the final output stages. Consequently, in practice it has been found that these noise signals may actually mask the information signals thereby causing the system to produce erroneous or null results. Hence, these systems are inherently subject to unreliable and inaccurate results.

Many solutions have been proposed to ameliorate the above problem but each solution has an inherent drawback. For example, one such solution is to use a narrow band receiver and heterodyne the received signal of unknown frequency with a local oscillator the frequency of which continuously varies. The theory behind this system is that at one point the received and local signals will mix to provide an intermediate signal which will pass through the receiver circuits. A similar system includes a relatively narrow bandpass filter in the receiver chain, the passband of which is varied. However, all of these systems are extremely inaccurate, particularly when they are utilized in acquisition radar schemes. Thus, since the system is scanning in both space and frequency tuning within the receiver, the signal indicating the target may be missed completely.

Accordingly, an object of the present invention is to provide an improved transmission system.

A more specific object of this aspect of the invention is to provide a transmission system which is specifically adapted for use as a radar acquisition or tracking system.

Another object of the invention is the provision of a transmission system which permits the continuous seeking and/or monitoring of a target regardless of Doppler shifts.

Another object of the invention is the provision of a transmission system which substantially eliminates the need for a stable local oscillator or the frequency scanning of the same or other circuit elements thereby reducing both the cost and the complexity of the associated circuitry.

A further disadvantage associated with presently known radar systems of the type described above arises even in those cases where the velocity of the target is known. To be more specific, as the radar system sweeps through an angle as it tracks a target the component of the velocity detected by the radar system continuously changes. Thus, the frequency shift of the reflected signal due to the Doppler effect likewise continuously varies and may fall out of the receiver passband. Such systems are designated angle dependent systems.

Accordingly, another object of the present invention is the provision of a transmission system which is angle independent even when utilized as a tracking radar system.

The Doppler effect occurs whenever one body is moving with respect to another. Thus, signals passing between a transmitter and a receiver will be shifted in frequency if either one or both elements of the system are in motion. Thus, in rocketship-to-rocketship communication systems, for example, the signals may be so shifted in frequency due to the velocities of the rocketships that the received signal may fall outside the passband of the receiver. The receiver may, of course, be tuned so that it will receive a wide band of frequencies; however, this solution produces problems similar to those described above with respect to widening the bandwidth of a radar receiver.

Accordingly, a further object of the invention is to provide a transmission system which insures the accurate retrieval of information from a transmitted signal regardless of movement of the transmitter and/or the receiver.

A more specific object of this aspect of the invention is to provide a high efficient and effective communication system which is operable to extract information modulated on a transmitted signal regardless of Doppler frequency shifts which occur in the system.

Accordingly, a transmission system constructed according to one aspect of the invention comprises transmission means for transmitting first and second frequency signals which are separated by a difference frequency whereby said first and second frequency signals may be received as respective third and fourth frequency signals which are separated substantially by the difference frequency. A receiver is provided which includes mixing means for mixing the third and fourth signals with a nonvaried fifth frequency signal to produce a plurality of mixed signals at least some of which are separated by the difference frequency. Filter means is connected to the mixing means for passing preselected ones of the plurality of mixed signals which are separated by the difference frequency. Detecting means is responsive to the preselected ones of the mixed signals for producing an output signal having a frequency related to the difference frequency. The nonvaried fifth frequency signal is preferably produced by a strong local oscillator.

In accordance with another aspect of the invention, the receiving system may be used separately for the improved passive emission detection of weak signals either emitted spontaneously (or stimulated by exposure to light, electrons, or other waves or particles) by an atom, molecule, collection of atoms, collection of mol-

ecules, liquid, gas, solid, or plasma. In this case, the two or more frequencies emitted and their difference frequency or frequencies may be unique to the source of radiation (which acts as the transmitter) enabling the final bandpass filter in the receiver to be tuned to a naturally-occurring difference frequency, providing near-optimum detectability for the species being detected. Examples of use in this mode include the determination of the existence and the measurement of distant (including atmospheric and extraterrestrial) species, including pollutants. Using the receiving system, detectability is unaffected by the Doppler shift caused by the gross and relative motion of these species. In the prior system, the unknown Doppler shifts due to both the gross and relative motion of such species (e.g. galactic nebulae or smokestack effluent) generally reduced the detectability enormously.

Other features and advantages of the present invention will become more apparent from a consideration of the following detailed description when taken in conjunction with the accompanying drawing, in which:

FIG. 1 is a schematic circuit diagram, in block form, of a transmission system constructed according to the present invention and whose receiving portion may be used separately for passive emission detection; and

FIG. 2 is a schematic circuit diagram, in block form, of a modified embodiment of a transmission system constructed according to the present invention wherein modulated information may be transmitted from a transmitter to a receiver.

In one sense, the system of the present invention is operable to transmit two signals which are separated by a difference frequency to a target. The two signals which are reflected or scattered from the target and which may be shifted in frequency due to Doppler effects are received by a receiving apparatus which is operable to extract a signal having the difference frequency, or substantially the difference frequency depending on the velocity of the target, from the received signals. Hence, the presence of this extracted signal indicates that the target is, in fact, present and may be continuously monitored.

In another sense the receiving system may be used separately for passive emission detection.

More specifically, a transmission system constructed according to the present invention is designated generally by the reference numeral **10** in FIG. **1** and includes a transmitter **12**. The transmitter **12** is a so-called two-frequency transmitter and transmits two signals at frequencies f_1 and f_2. (For ease of reference, these signals will be referred to as signals f_1 and f_2 and the reflected signals, which are frequencies f_3 and f_4, will similarly be referred to as signals f_3 and f_4.) The transmitted signals f_1 and f_2 are separated by a difference frequency Δf. The signals f_1 and f_2 may be optical, infrared, or microwave frequency signals and the only design limitation on the transmitter **12** is that the difference frequency Δf remains substantially constant albeit the actual frequency of the signals may drift somewhat. This limitation is easily achieved if the transmitter is a two-mode laser since the modes tend to drift together thereby keeping the difference frequency Δf constant. A practical laser which has these properties is a carbon dioxide laser. However, this example is not to be considered as being a limitation of the present invention since any type of transmitter which maintains the difference frequency substantially constant may be utilized.

As noted above, the signals f_1 and f_2 are transmitted to a target **14**. Under normal circumstances, the target **14** which may, for example, be a satellite, an airplane, an astronomical body, or the like is moving and the signals impinging on the target **14** and being reflected or scattered therefrom will be shifted in frequency due to Doppler effects. In addition to frequency shifts due to the velocity of the target, it should be noted that the signals may also be broadened in frequency due to the scattering of the signals. Accordingly, after reflection or scattering, the signal f_1 will have a frequency f_3 and the signal f_2 after reflection or scattering will have the frequency f_4. However, as noted in greater detail below, the difference in frequency between the signals f_3 and f_4 will be substantially equal to the difference frequency Δf.

The signals f_3 and f_4 are received by a receiving apparatus designated generally by the reference numeral **16**. The receiving apparatus or receiver **16** includes a mixer or mixing stage **18** to which the signals f_3 and f_4 are applied. It is to be understood that the stage **18** may include a photodetector or other detector of electromagnetic radiation as well as amplifiers for amplifying the received signals. A local oscillator **20** applies a signal f_L to the mixer **18**. The mixer **18** is operable to mix three signals together. That is, the mixer **18** is operable to mix the signal f_L from the local oscillator **20** with the received signals f_3 and f_4 to produce a plurality of signals which may include sum and difference frequency signals as well as intermodulation signals.

In practice, where a carbon dioxide or other infrared or optical laser is used as the source of the signals, the mixer will usually produce only difference frequency components and a dc component and not sum frequency and double-frequency components. Additionally, it is desirable to provide a relatively strong local oscillator signal f_L so that the signal-to-noise ratio is maximized, in which case the signal of frequency f_3-f_4 will be relatively weak. However, these examples are for illustrative purposes only and are not to be interpreted as being a limitation on the present invention.

The output terminals of the mixer **18** are connected to a bandpass filter **22** through a blocking or coupling capacitor **24** which blocks the DC component of the mixed signals produced by the mixer **18**.

The passband of the filter **22** is chosen so that only the difference frequencies in the band of frequencies under consideration will appear at the output terminals thereof. That is, the passband of filter **22** is selected to be narrow to reduce the amount of noise which may pass through the filter. However, the filter, although limiting noise, still is of sufficient width to pass the difference signals produced by the mixer **18**. Hence, in the example under consideration, the signals appearing at the output terminals of the bandpass filter will comprise a signal having the frequency $(f_3 - f_L)$ and a signal having the frequency $(f_4 - f_L)$. Since the signals f_3 and f_4 were substantially separated by the difference frequency Δf, it is obvious that the signals $(f_3 - f_L)$ and $(f_4 - f_L)$ will similarly be substantially separated by the difference frequency Δf.

The signals appearing at the output terminals of the bandpass filter **22** are applied to a non-linear device **26**. The non-linear device **26** is operable to produce a signal having a component at the difference frequency Δf. That is, the cross product of the input signals $(f_3 - f_L)$ and $(f_4 - f_L)$ will produce a signal at the output whose

peak is centered at substantially the frequency Δf. One such non-linear device may be a square-law device. However, it is emphasized that this is by way of illustration only and is not to be interpreted as being a limitation on the present invention. That is, similar results would be expected with any linear device such as a non-linear linear device in place of the square-law device.

The output terminals of the non-linear device **26** are connected to a bandpass filter **28** which has a relatively narrow passband centered near or about the difference frequency Δf. Accordingly, the signal appearing on a lead **30** which is connected to the output terminals of the filter **28** will have a frequency very close or equal to Δf. This signal may then be applied to appropriate indicating or processing circuitry to indicate the presence of the detected signal.

Summarizing the operation of the above system, the transmitted signals which are substantially separated by a difference frequency Δf are received by the receiver **16** and are heterodyned by a mixer **18** in conjunction with a local oscillator **20** to produce difference signals. These difference signals are applied to a non-linear device which produces a signal at substantially the frequency Δ f. The signals are applied to a bandpass filter which passes the signal having substantially the frequency Δ f and applies the same to appropriate indicating circuitry. Accordingly, a system has been described which permits facilitated acquisition and continuous monitoring of a target and which eliminates the need for high-frequency electronics and their attendant disadvantages, in view of the fact that a low-frequency signal very close to Δ f is processed by indicating circuitry and the like rather than higher frequency signals. This system provides an ease in impedance matching as well as a high signal-to-noise ratio and a low minimum-detectable-power.

Alternatively, the receiving apparatus **16** may be used separately for passive emission detection.

Another advantage of the present system is that the bandwidth of the system can be extended over a greater range than conventional radar systems. That is, the final indicating circuit is still tuned to substantially the frequency Δf even though the bandwidth of the bandpass filter **22** may be increased. Additionally, the present system eliminates the need for the electronic tuning of frequency sensitive circuits and their attendant cost and complexity.

As noted above, it has been assumed that the frequency difference between signals f_3 and f_4 is substantially equal to the frequency difference Δf between signals f_1 and f_2. To put this another way, it has been assumed that the frequency difference Δf after undergoing a Doppler shift (i.e., the difference between signals f_3 and f_4) is substantially equal to the difference Δf before such Doppler shift (i.e., the difference between signals f_1 and f_2). To put this still another way, it has been assumed that the difference between the Doppler shifted difference signal and the difference signal before Doppler shift is zero. Thus, consider an example utilizing a carbon dioxide laser radar operating at 10.6 mm in the infrared region. Assuming that a difference frequency Δf is chosen with a value of 1 MHz and that the target is a satellite having a 1 meter radius with a rotation rate of 1 rpm, and a radial velocity of approximately 10 km/sec; then, the Doppler frequency will be equal to approximately 2 GHz. Utilizing these quantities, calculations show that the difference between the Doppler shifted difference frequencies f_3 and f_4 and the difference Δf between the transmitted signals f_1 and f_2 is approximately equal to 60 Hz. This is a very small shift and justifies the assumption that the difference frequency between signals f_3 and f_4 is substantially equal to the difference frequency between signals f_1 and f_2. (For more detailed analysis of this calculation, the reader is referred to the article: Three-Frequency Heterodyne System for Acquisition and Tracking of Radar and Communications Signals, by the inventor, which appears in Volume 15, Number 12 of the *Applied Physics Letters* of Dec. 15, 1969, pages 420–423, hereby incorporated by reference.)

The transmission system of FIG. **1** may also be utilized to eliminate clutter from the received signal. That is, where it is desired to seek or track a moving target, stationary objects may produce signals or "clutter" which may mask the desired signals. Accordingly, in order to eliminate such clutter a notch filter **31** is provided which is connected to the output terminals of the bandpass filter **28.** The output terminals of the filter **31** are connected to the appropriate indicating circuitry. The notch filter is tuned precisely to the known difference frequency Δf and has an extremely sharp characteristic so that only signals at the frequency Δf will be attenuated by the filter.

In operation, if the radar signal is reflected or scattered from a stationary object it will not undergo any Doppler shift and, accordingly, the filter **28** output signal will be exactly equal to Δf. However, this signal will be attenuated by the filter **31.** On the other hand, if the signal is reflected or scattered from a moving target, the signal appearing at the output terminals of the bandpass filter **28** will be nearly but not exactly equal in frequency to Δf. Thus, this latter signal will pass through the notch filter **31** to indicate the moving target.

Accordingly, the transmission system of the present invention accurately indicates a moving target in the presence of stationary objects.

In conventional communication systems wherein information modulated on a signal is transmitted to a receiver, the transmitted signal may be Doppler shifted in frequency due to the movement of either the transmitter, the receiver, or both. Additionally, the signal may be shifted in frequency due to reflection or scattering of the signal from a moving target. Thus, if the shifted frequency signal falls outside the passband of the receiver, the information will be lost. Accordingly, FIG. **2** illustrates a transmission system constructed according to the present invention which may be utilized as a communications system to transmit information from a transmitter to a receiver in the presence of Doppler shifts.

While the embodiment of FIG. **2** illustrates the more complex case wherein the Doppler shift is produced by reflecting or scattering the signal from a moving target as well as from motion of the transmitter and/or receiver, it is to be noted that this is for illustrative purposes only. That is, in the simpler case the transmitter transmits directly to the receiver and the Doppler shift occurs only because of relative movement therebetween.

More specifically, the communications system of FIG. **2** is designated generally by the reference numeral **32** and includes transmitting apparatus **34.** The appara-

tus 34 is adapted to transmit signals having frequencies f_1 and f_2 which are separated by a difference frequency Δf. The transmitting apparatus may include a two-frequency laser of the type noted above or, alternatively, two single-frequency lasers. Additionally, one of the frequency signals produced by the transmitting apparatus 34 may be modulated in accordance with known modulating techniques so that, for example, the signal f_2 will carry information modulated thereon.

Thus, as shown in FIG. 2, a source 33 of signals f_1 is provided. Additionally, a source 35 of signals f_2 is provided which is modulated by a modulator 37. That is, only one of the transmitted signals is modulated.

The signals f_1 and f_2 may be reflected from a mirror 36 to a moving target 38. Of course, if transmission is directly to the receiver the mirror 36 may be eliminated. The signal reflected or scattered from the moving target 38 will be shifted in frequency due to Doppler shift and, accordingly, the signal f_1 will be returned as signal f_3, and the modulated signal f_2 will be returned as signal f_4. In other words, the signal of frequency f_3 will be the transmitted signal f_1 but shifted in frequency due to the Doppler shift and the signal of frequency f_4 will be the signal f_2 also shifted in frequency due to Doppler effect. The returned signals f_3 and f_4 are received by a receiver designated generally by the reference numeral 40.

The receiver 40 includes a beamsplitter 42 through which the signals f_3 and f_4 pass to the photodetector and mixer 44. Associated with the mixer 44 is a local oscillator 46 which may comprise a single-frequency laser which produces a beam at a frequency f_L. The output signal of the oscillator 46 is reflected from the beamsplitter 42 to the photodetector and mixer 44. The photodetector and mixer 44, similarly to the mixer 18, is operable to produce the sum, double, and difference frequency signals between f_3, f_4 and f_L and intermodulation signals. The remainder of the receiving chain in the receiver 40 is similar to the chain in the receiver 16. That is, connected to the mixer 44 is a series circuit comprising the capacitor 24, bandpass filter 22, non-linear device 26 and a bandpass filter 28 which is centered at the difference frequency Δf. The output signal from the bandpass filter 28 is applied to a demodulator 48 which demodulates the information carried by the signal of center frequency substantially equal to Δf so that the information impressed on the original waveform may be retrieved.

The operation of the system of FIG. 2 is similar to the operation of the system of FIG. 1 described above, with the exception that the signal f_2 is modulated, as noted above. The non-linear device 26 in the apparatus of FIG. 2 produces a component centered substantially at the frequency Δf. This frequency signal is applied through the bandpass filter 28 to the demodulator 48. By modulating only one of the original signals f_2, the signal by signal component reaching the demodulator 48 through the bandpass filter 28 results from the convolution of a delta-function (at frequency f_1) with the modulated signal (centered at f_2), and is simply the original modulated information.

Accordingly, the above described communications system is operable to directly retrieve information from a modulated signal regardless of frequency shifts due to Doppler effects.

It is also to be noted that the systems of FIG. 1 and FIG. 2 also produce accurate results in the presence of scattering or atmospheric effects which, in effect, produce a frequency broadening of each transmitted signal. Moreover, the receiving portion of the system of FIG. 1 may be used separately for passive emission detection.

While preferred embodiments of the invention have been shown and disclosed herein, it will be obvious that numerous omissions, changes and additions may be made in such embodiments without departing from the spirit and scope of the present invention.

What is claimed is:

1. A transmission system comprising transmission means for transmitting first and second frequency signals separated by a difference frequency to a moving target whereby said first and second frequency signals may be received as respective third and fourth frequency signals which differ in frequency from said first and second frequency signals because of the Doppler effect but which are separated by substantially said difference frequency, and receiving means for receiving said third and fourth frequency signals, said receiving means comprising local signal generation means for generating a local signal at a nonvaried fifth frequency, mixing means for mixing said third and fourth frequency signals with said fifth frequency signal to produce a plurality of mixed signals including signals which are separated by substantially said difference frequency, filter means connected to said mixing means for passing preselected ones of said plurality of mixed signals which are separated by substantially said difference frequency, and detecting means responsive to said preselected ones of said mixed signals for producing an output signal having a frequency related to said difference frequency.

2. A transmission system as in claim 1, in which said transmission means is a two-mode laser.

3. A transmission system as in claim 2, in which said laser is a carbon dioxide laser operating at said first and second frequencies.

4. A transmission system comprising transmission means for transmitting first and second frequency signals separated by a difference frequency to a moving target whereby saiad first and second frequency signals may be received as respective third and fourth frequency signals which differ in frequency from said first and second frequency signals becuse of the Doppler effect but which are separated by substantially said difference frequency, and receiving means for receiving said third and fourth frequency signals, said receiving means comprising local signal generation means for generating a local signal at a fifth frequency, which local signal is substantially stronger than said received third and fourth frequency signals, mixing means for mixing said third and fourth frequency signals with said substantially stronger fifth frequency signal to produce a plurality of mixed signals including signals which are separated by substantially said difference frequency and are substantially stronger than signals produced by mixing together said third and fourth frequency signals, filter means connected to said mixing means for passing preselected ones of said plurality of mixed signals which are separated by substantially said difference frequency, and detecting means principally responsive to the substantially stronger of said preselected ones of said mixed signals for producing an output signal having a frequency related to said difference frequency.

5. A transmission system as in claim 4, in which said filter means comprises a bandpass filter having a passband which encompasses signals having frequencies equal to said fourth frequency minus said fifth frequency and said third frequency minus said fifth frequency.

6. A transmission system as in claim 5, in which said detecting means comprises a non-linear device response to said preselected ones of said mixed signals for producing a signal having a frequency substantially equal to said difference frequency, and a filter connected to said non-linear device for passing said signal having a frequency substantially equal to said difference frequency.

7. A transmission system as in claim 6, in which said non-linear device comprises a square-law device.

8. A transmission system as in claim 1, in which said transmission means includes a modulator for modulating said first frequency signal with preselected information, and said detecting means comprises a non-linear device responsive to said preselected ones of said mixed signals for producing a signal having a frequency substantially equal to said difference frequency, and having said information modulated thereon, a filter connected to said non-linear device having a passband which encompasses said signal produced by said non-linear device, and a demodulator connected to said filter for demodulating the signal passed by said filter.

9. A transmission system as in claim 1, and filter means connected to said detecting means for attenuating those signals having a frequency precisely equal to the difference in frequency between said first and second frequency signals.

10. A transmission system as in claim 9, in which said filter means comprises a notch filter.

11. A transmission system comprising a transmitter for transmitting first and second frequency signals separated by a first difference frequency to a moving target and a receiver for receiving third and fourth signals respectively related to said first and second signals and separated by a second difference frequency, said receiver comprising heterodyne means for mixing a fifth frequency signal with said signals to produce a plurality of mixed signals, a first filter for passing preselected ones of said plurality of mixed signals which are separated by said first and second difference frequencies, a non-linear device responsive to said preselected ones of said signals for producing an indication signal having a frequency exactly equal to said second difference frequency, and a second filter for attenuating a signal having a frequency exactly equal to said first difference frequency and for passing only said indication signal.

12. A transmission system as in claim 11, in which said non-linear device is a square-law device.

13. A transmission system as in claim 11, in which said transmitter is a two-mode laser.

14. A transmission system as in claim 6, in which said transmission means comprises a modulator for modulating at least said first frequency signal with preselected information, and said receiving means comprises a demodulator connected to said second filter for demodulating the signal passed thereby.

15. A transmission system comprising transmission means for transmitting first and second frequency signals separated by a first difference frequency to a moving to a moving target whereby said first and second frequency signals may be received as respective third and fourth frequency signals which are separated by a second difference frequency which is slightly different from said first difference frequency, and receiving means for receiving said signals, said receiving means comprising means for detecting a signal only having a frequency exactly equal to said second difference frequency.

16. A signal detection system for detecting a signal source which produces first and second frequency signals separated by a first difference frequency which are received at a signal receiving position as respective third and fourth frequency signals which differ in frequency from said first and second frequency signals because of Doppler effect caused by relative motion between the signal source and signal detecting position and which are separated by a second difference frequency substantially equal to said first difference frequency, said signal detection system comprising receiving means at said signal receiving position for receiving said third and fourth frequency signals, local signal generation means for generating a local signal at a nonvaried fifth frequency, mixing means for mixing said third and fourth frequency signals with said fifth frequency signal to produce a plurality of mixed signals including mixed signals which are separated by substantially said second difference frequency, non-linear means responsive to said mixing means for producing signals including a signal having a frequency equal to said second difference frequency, filter means, responsive to said non-linear means, having a bandpass including said second difference frequency for passing a signal having a frequency equal to said second difference frequency, and indicating means responsive to said filter means for producing an output signal indicating the detection of said signal source.

17. A signal detection system for detecting remote or local multiple-signal radiating matter at a signal detection position when there is relative motion between the multiple-signal radiating matter and the signal detection position, said multiple-signal radiating matter radiating at least two signals having first and second frequencies together with a first difference frequency signal having a frequency equal to the difference between such first and second frequencies, said signal detection system comprising:

A. receiving means for receiving third and fourth frequency signals which differ in frequency from said first and second frequency signals because of the Doppler effect which are separated by a second difference frequency which is substantially equal to said first difference frequency;

B. local signal generation means for generating a local signal at a nonvaried fifth frequency;

C. mixing means for mixing said third and fourth frequency signals with said fifth frequency signal to produce a plurality of mixed signals including mixed signals which are separated by substantially said second difference frequency;

D. non-linear means responsive to said mixing means for producing signals including an output signal having a frequency equal to said second difference frequency;

E. filter means connected to said non-linear means for passing only the signals having a frequency substantially equal to said second difference frequency; and

F. indicating means responsive to said filter means for producing an output signal indicating the detection of said multiple-signal radiating matter.

18. The signal detection system of claim 17 wherein said local signal generation means generates a non-varied fifth frequency signal which is substantially stronger than said received third and fourth frequency signals.

19. The signal detection system of claim 18 wherein said mixing means produces a plurality of mixed signals including signals which are separated by substantially said difference frequency and are substantially stronger than signals produced by mixing said third and fourth frequency signals.

20. The signal detection system of claim 17 including further filter means connected between said mixing means and said non-linear means having a passband which encompasses signals having frequencies equal to said fourth frequency minus said fifth frequency and said third frequency minus said fifth frequency.

21. The signal detection system of claim 19 further comprising an attenuating means responsive to said filter means for attenuating signals having a frequency precisely equal to said first difference frequency.

22. The signal detection system of claim 21 wherein said attenuating means comprises a notch filter.

* * * * *

United States Patent [19]

Berrod et al.

[11] 3,942,181

[45] Mar. 2, 1976

[54] VARIABLE-GAIN AMPLIFIER

[75] Inventors: François Berrod; Fernand Puverel, both of Paris, France

[73] Assignee: Thomson-CSF, Paris, France

[22] Filed: Oct. 19, 1973

[21] Appl. No.: 407,996

[30] Foreign Application Priority Data

Oct. 20, 1972 France 72.37263

[52] U.S. Cl. .... 343/16 M; 330/28; 330/86; 330/29; 330/145

[51] Int. Cl.² G01S 9/22; H03F 1/34

[58] Field of Search......... 330/29; 145, 28, 144, 86, 330/85; 343/16 M

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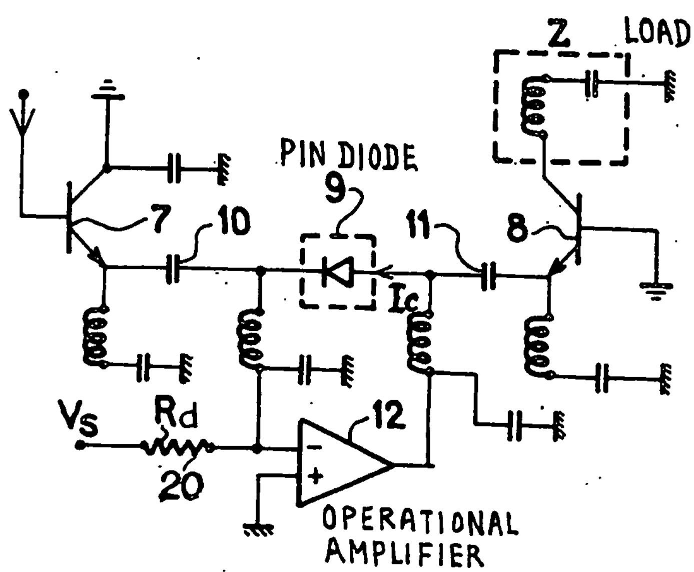
Primary Examiner—T. H. Tubbesing
Assistant Examiner—G. E. Montone
Attorney, Agent, or Firm—Karl F. Ross; Herbert Dubno

[57] ABSTRACT

A variable-gain amplifier comprises a transistor with an emitter connected in a degenerative emitter-follower circuit which includes a PIN diode traversed by a current injector in the form of an operational amplifier, the PIN diode forming part of a negative-feedback path extending from the output to the inverting input of the operational amplifier. To achieve a gain characteristic whose logarithm varies linearly with a control voltage, this control voltage is fed to the same inverting input through an amplifier circuit with an exponential characteristic, including another operational amplifier provided with an input transistor. A temperature-compensating circuit, preceding that input transistor, comprises still another operational amplifier having a further transistor inserted in its negative-feedback path. The variable-gain amplifier may be included in one of two conjugate (sum and difference) channels of a monopulse radar.

10 Claims, 5 Drawing Figures



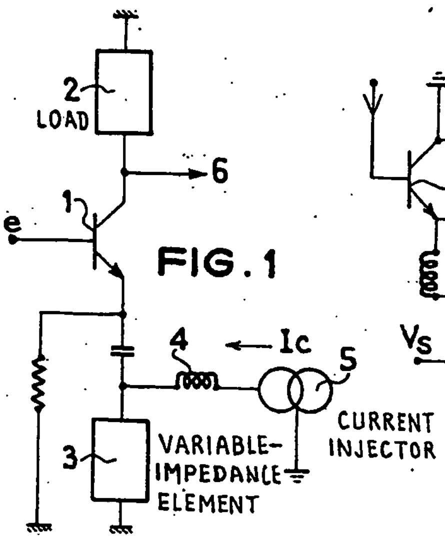
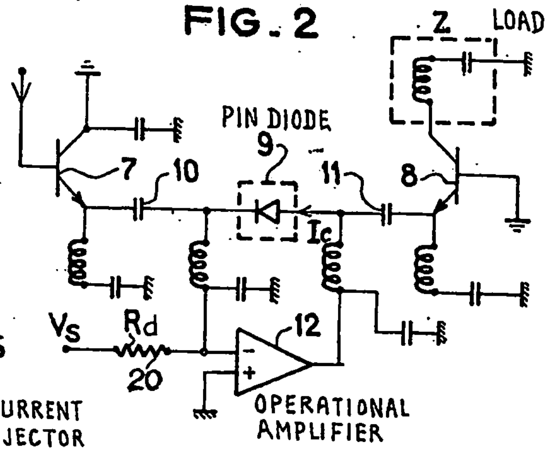


FIG. 1

FIG. 2



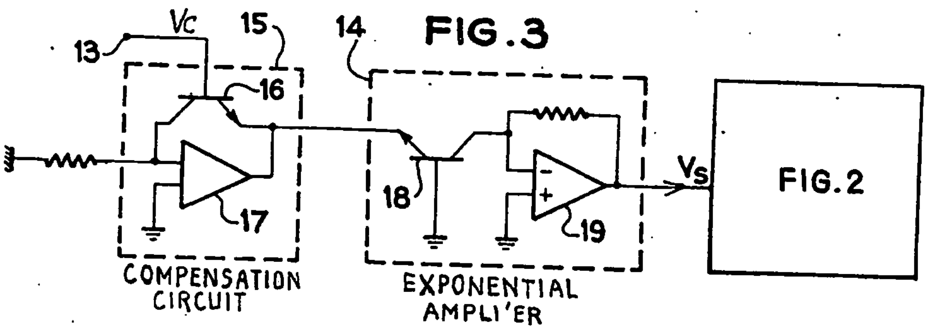


FIG. 3

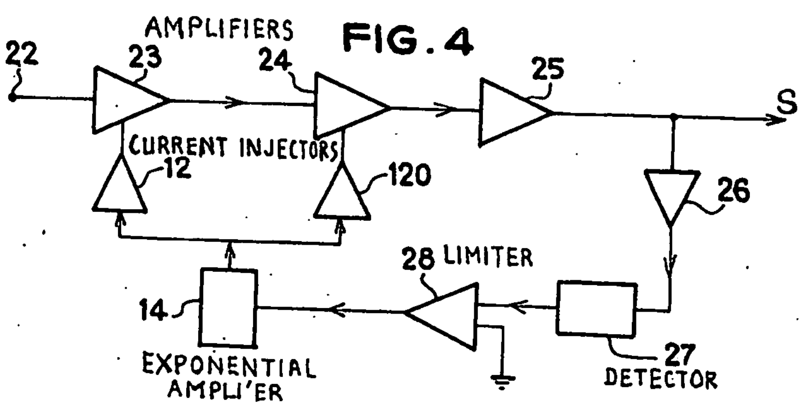


FIG. 4

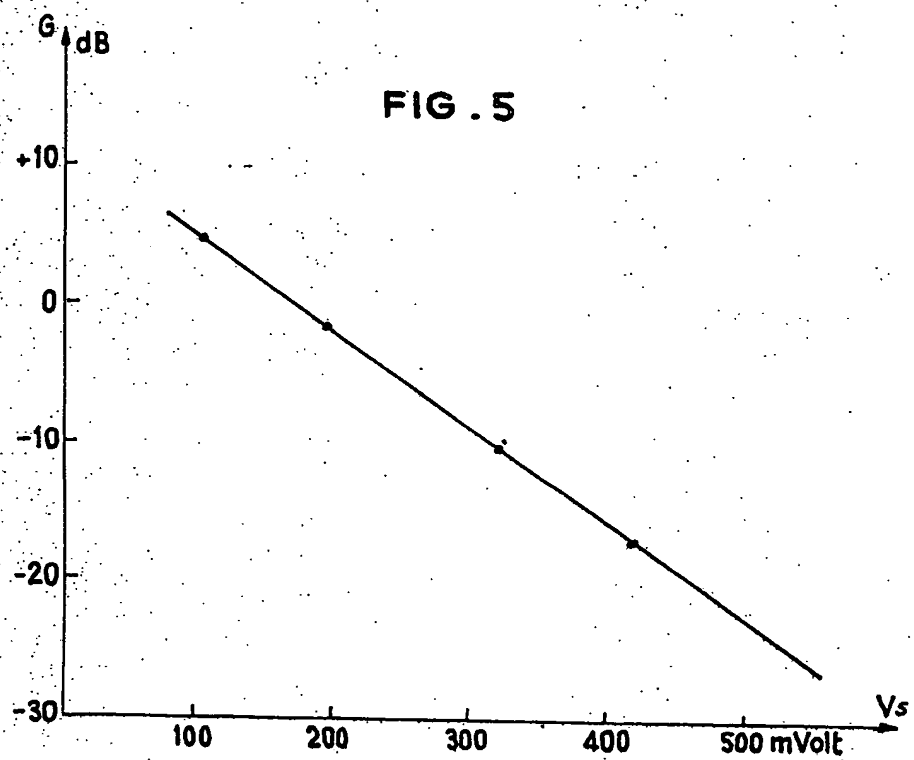


FIG. 5

VARIABLE-GAIN AMPLIFIER

BACKGROUND OF THE INVENTION

The present invention relates to improvements in variable-gain amplifiers requiring gain control varying in accordance with a well defined law, e.g., linearly, over a wide dynamic range.

Prior attempts at solving this problem have not turned out satisfactory. Thus, a gain control has been proposed including an attenuator consisting of a fixed impedance and a nonlinear element such as a diode, whose dynamic impedance varies as a function of the direct current flowing through it. Another device utilizes field-effect transistors whose gate voltage is varied. Yet another device of this type comprises a variable-resistance element of the diode type, arranged in a negative-feedback path of the variable-gain amplifier to be controlled.

However, these devices have drawbacks. In particular, the gain varies as a function of the temperature and this militates against good reproducibility. Moreover, the law governing the variation of the logarithm of the gain as a function of the control voltage is not linear and the dynamic range is relatively narrow, thereby severely limiting the usefulness of the controlled amplifier stage.

SUMMARY OF THE INVENTION

The object of the present invention is to overcome these drawbacks and to provide a gain control of wide dynamic range for an amplifier stage, the gain varying in accordance with a readily reproducible characteristic. More specifically, our invention aims at providing a gain characteristic whose logarithm varies linearly with the control voltage.

In accordance with the present invention, a variable-impedance element constituted by a PIN diode is connected in a degenerative or negative-feedback circuit connected to a reference electrode (e.g., emitter) of a transistor, or pair of transistors, included in the controlled amplifier stage, the PIN diode being biased by a current flowing in a negative-feedback path of an operational amplifier supplied at an input thereof with the gain-controlling voltage.

BRIEF DESCRIPTION OF THE DRAWING

The above and other features of our invention will now be described in detail with reference to the accompanying drawing wherein:

FIG. 1 is a circuit diagram of a variable-gain amplifier embodying our invention;

FIG. 2 is a circuit diagram of another embodiment;

FIG. 3 illustrates additional components of a system including the variable-gain amplifier of FIG. 2;

FIG. 4 is a block diagram illustrating the application of our improved variable-gain amplifier to a monopulse radar system; and

FIG. 5 is a graph showing the variation of the gain G of the amplifier of FIG. 4 as a function of the applied control voltage.

DETAILED DESCRIPTION

FIG. 1 illustrates schematically an amplifier, or amplifier stage, comprising an element 3 whose impedance varies as a function of the current flowing through it, connected in a negative-feedback circuit in the amplifier stage. This stage, in a manner known per

se, comprises a transistor with an emitter-follower connection. The collector of the transistor 1 is connected to a load impedance 2 across the terminals of which the output signal is picked up at **6** in response to an input signal *e* applied to the base. The emitter of the transistor 1 is grounded for high frequencies through a capacitor in series with the variable-impedance element 3 which is connected, through a surge coil **4**, to a current injector 5 which supplies the biasing current *Ic* for the element 3.

The element **3**, whose resistance varies with the current flowing through it, is a PIN (positive/intrinsic/-negative diode.

The characteristics of these diodes are well known, and need only be outlined briefly here. A PIN diode consists of two heavily doped regions, p+ and n+, known as the end regions, separated by a lightly doped intermediate region. When the intermediate region has a substantial thickness (on the order of 10 to 100 microns), the diode acts as a high-voltage rectifier having a low direct-voltage drop for high currents because of the modulation of the conductivity of the intermediate region by the high number of charge carriers injected through the end regions. However, in the very-high-frequency range, a PIN diode of this kind can function as a variable resistor because the frequencies are then too high for the rectification to take place on account of the relatively long recovery time of the intermediate layer.

For zero or reverse bias, the intermediate layer creates a high resistance. Under the effect of forward bias, the injection and storage of charge carriers reduces this resistance in the intermediate region in accordance with the formula

$$R \approx \frac{(20\text{–}50)}{I^{0.87}}$$

where R is expressed in ohms and *I* is the forward biasing current expressed in milliamps.

It is in this latter state that the PIN diode is utilized in the system embodying our present invention, where it affords numerous advantages.

The insertion of the variable element in a feedback circuit makes it possible, first of all, to achieve a dynamic range in the variable-gain stage which increases with the input level. In the negative-feedback amplifier, the dynamic-gain variation is limited by the impedance of the electrode to which the variable element is coupled, the electrode in this case being the emitter of the transistor, on the one hand, and by the parasitic capacitances of the circuit, on the other.

For a PIN diode, the charge-carrier life greatly exceeds the cycle length corresponding to the frequency of the signals applied to the amplifier. Thus, by using the PIN diode as a variable element, complete separation between the dynamic and static characteristics of the diode is achieved, that is to say between the dynamic impedance of the diode and its low-frequency impedance. Consequently, with a charge-carrier life on the order of 1.3 microsecond, there is a lower limit on the order of 1 megahertz for the frequency at which the diode will operate as a variable resistor. Likewise, there is an upper limit for the frequency of the diode-biasing current. This makes it possible to achieve very low-distortion amplification since the permissible dynamic current becomes independent of the control current.

Moreover, by connecting the PIN diode in a negative-feedback circuit including a current injector **5**, we are able to insure that the current flowing through the diode depends only upon the input voltage f the current injector and this improves the gain control of the amplifier, making it possible to effect better utilization of the wide control range afforded by the emitter-follower connection.

FIG. **2** illustrates another embodiment of the invention, in which the controlled amplifier comprises a differential circuit.

This two-stage amplifier, constructed as an integrated circuit, comprises two transistors **7** and **8** of like conductivity type (NPN) in emitter-follower connection with their emitters interconnected. The element whose resistance varies with the current flowing through it, i.e., the PIN diode **9** is connected in the link between the emitters of the transistors **7** and **8**, through the intermediary of respective capacitors **10** and **11**; each emitter, moreover, is connected to ground through a surge coil and a decoupling capacitor. It will be observed, too, that the signals processed in this amplifier stage are applied to the base of the transistor **7**, the output signals from the amplifier stage being picked up at the collector of the transistor **8** across the terminals of the load Z. The PIN diode **9**, connected in a coupling network included in the degenerative emitter-follower circuit of input transistor **7** and inserted between the emitters of the amplifiers **7**, **8** also forms part of a negative-feedback circuit associated with an operational amplifier **12** included in the gain control circuit of the amplifier.

The operational amplifier **12** converts a control voltage V*s*, applied to its inverting input through a resistor **20**, into a biasing current I*c* acting directly on the PIN diode **9**.

The system of FIG. **2** has the advantage, in relation to one with a simple amplifier of the kind shown in FIG. **1**, of still further increasing the dynamic range of the signal which the amplifier will be able to handle.

Accordingly, the amplitude of the permissible signal at the amplifier input, that is to say at the base of the transistor **7**, increases as the impedance of the diode **9** increases, that is to say as the gain decreases. The variation in gain which it is thus possible to achieve has no effect upon the pass-band characteristics of the amplifier stage, which is determined by the load Z separated from the PIN diode **9** by the transistor **8**. As already mentioned, this gain variation is independent of active elements other than the PIN diode **9**, the stage gain being given by the ratio of the load resistance Z connected to the collector of the transistor **8**, to the dynamic resistance of the PIN diode.

Still better conditions of operation of the controlled amplifier can be obtained in accordance with the invention by modifying the control voltage applied to the input of the operational amplifier **12** which directly controls the current I*c* flowing through the PIN diode.

In the introduction to the present specification we have pointed out that the logarithm of the gain of the amplifier stage should vary as a linear function of the control voltage.

This condition is met in accordance with the invention by connecting between a terminal **13** (FIG. **3**), receiving the control voltage V*c* which determines the gain of the amplifier **7**, **8** of FIG. **2**, and the operational amplifier **12**, acting directly upon the variable-resistance element, an amplifier circuit **14** with an exponential response.

The control voltage V*c* appearing at **13** is applied to the exponential amplifier circuit **14** through a compensation circuit **15** comprising a transistor **16** connected in a feedback path between the inverting input and the output of an operational amplifier **17**. This latter amplifier, which is of a type well known per se, requires no detailed description, any more than does the operational amplifier **12**. The circuit **15** is designed to compensate for the variation in the base-emitter voltage of the transistor **16** as a function of temperature. The current flowing through a grounded-base input transistor **18** of circuit **14**, whose collector is connected to inverting input of an operational amplifier **19** thereof, thus varies exponentially with the base-emitter voltage of input transistor **18** which is connected in series-opposed relationship with feedback transistor **16**. Consequently, the voltage V*s* appearing at the input of the resistor **20** (FIG. **2**), connected to the operational amplifier **12**, is exponentially related to the voltage V*c*; the current I*c* flowing through the PIN diode **9**, which is equal to the quotient of the value V*s* divided by the magnitude R*d* of resistor **20**, varies exponentially with the voltage V*c* so that the logarithm of the gain is a linear function of V*c*. *voltage*

Thus, we have disclosed a variable-gain amplifier controlled by a PIN diode connected to an emitter of a transistor amplifier. Alternatively, the diode could be connected in the collector circuit, without sacrificing the advantages of our invention.

The variable-gain amplifier in accordance with the invention has numerous applications, for example in any controlled-gain amplifier system operating within a frequency range extending from 1MHz to 1GHz, with this both closed-loop operation, where an automatic gain control (AGC) is created, and open-loop operation, using for example time-variable gain (TVG). A particularly significant application, however, is in a three-amplifier system for automatic gain control as applied to the sum and difference channels of a monopulse radar receiver using amplitude processing, in which the quality of exploration depends primarily upon the gains of the three amplifiers being identical throughout the gain-control range.

FIG. **4** schematically illustrates this kind of automatic-gain-control loop, in accordance with the invention, as applied to one of the conjugate (sum and difference) channels comprising two controlled-gain amplifier stages.

The channel in question, which may be the sum channel of a monopulse radar, thus comprises two controlled-gain amplifiers **23** and **24**, followed by an amplifier **25** delivering the output of the channel. The output signals are also applied to the gain-control loop of the amplifiers **23** and **24** through an amplifier **26** supplying a detector stage **27** connected to a threshold amplifier or limiter **28** in a circuit whose time constant determines the pass band of the control system. The amplifier circuit **28** is connected to the amplifier circuit **14** of FIG. **3**, having an exponential response characteristic, which in turn is connected to two current injectors **12** and **120**, that is to say operational amplifiers controlling PIN diodes (not shown here) arranged in a negative-feedback circuit f the controlled variable-gain amplifiers **23** and **24**. Identical injectors contr l the variable-gain amplifiers of the ther channel, i.e., of the difference channel.

FIG. 5 illustrates by way of example the attenuation law of the channel described with reference to FIG. 4, i.e., the variation in gain G as a linear function of the control voltage V_s.

By way of example, in closed-loop operation, the level may be regulated between 10 and 65 dB per mW in continuous-wave operation or in intermittent operation with pulses on the order of 0.25 microsecond.

It is clear, in the view of the excellent linearity of the impedance variation in the Pin diode as a function of the current flowing through it, that in accordance with the intended applications other laws of variation of the gain control are conceivable. Thus, a $\sin^2x/x^2$ law may be required if the amplifier is associated for example with a simulator.

What is claimed is:

1. A variable-gain amplifier comprising:
semiconductor means provided with an input electrode connected to a source of high-frequency signals and with two further electrodes including a reference electrode whose potential relative to that of said input electrode determines the conductivity of said semiconductor means, one of said further electrodes being connected to a load;
degenerative circuitry connected to said reference electrode for applying thereto a negative-feedback voltage;
a PIN diode in said degenerative circuitry;
biasing means for said PIN diode including an operational amplifier provided with an inverting input, a non-inverting input, an output, and a negative-feedback path connected between said output and inverting input thereof, said PIN diode being inserted in said negative-feedback path; and
a source of gain-controlling voltage connected to one of the inputs of said operational amplifier.

2. A variable-gain amplifier as defined in claim 1 wherein said semiconductor means comprises a first and a second transistor stage each having a base, an emitter and a collector, said input electrode being the base of said first transistor stage, said reference electrode being the emitter of said first transistor stage, said degenerative circuitry forming part of an emitter-follower connection for said first transistor stage and being coupled to the emitter of said second transistor stage, the emitter of said first transistor stage being connected to the load via the emitter and collector of said second transistor stage.

3. A variable-gain amplifier as defined in claim 2 wherein said transistor stages are of the same conductivity type.

4. A variable-gain amplifier as defined in claim 1 wherein said source of gain-controlling voltage comprises an amplifier circuit with an exponential characteristic.

5. A variable-gain amplfier as defined in claim 4 wherein said source of gain-controlling voltage further comprises a temperature-compensating circuit preceding said amplifier circuit.

6. A variable-gain amplifier as defined in claim 5 wherein said amplifier circuit and said temperature-compensating circuit comprise a second and a third operational amplifier, respectively.

7. A variable-gain amplifier as defined in claim 6 wherein said amplifier circuit includes an input transistor connected to said second operational amplifier, said temperature-compensating circuit including a further transistor inserted in a negative-feedback connection of said third operational amplifier, said further transistor being provided with a control-voltage input.

8. A variable-gain amplifier as defined in claim 7 wherein said input transistor and said further transistor are connected in series-opposed relationship.

9. In a monopulse radar having two conjugate channels, the improvement wherein one of said channels includes a variable-gain amplifier comprising:
semiconductor means provided with an input electrode connected to a source of high-frequency signal and with two further electrodes including a reference electrode whose potential relative to that of said input electrode determines the conductivity of said semiconductor means, one of said further electrodes being connected to a load;
degenerative circuitry connected to said reference electrode for applying thereto a negative-feedback voltage;
a PIN diode in said degenerative circuitry;
biasing means for said PIN diode including an operational amplifier provided with an inverting input, a noninverting input, and output, and a negative-feedback path connected between said output and inverting input thereof, said PIN diode being inserted in said negative-feedback path; and
a source of gain-controlling voltage connected to one of the inputs of said operational amplifier.

10. The programmable controller as recited in claim 9 in which said scanner circuit includes control means for alternately executing a data output sequence and a data input sequence, said control means being connected to said interrupt means, connected to said data in gate means and connected to said data out gate means, and being operable to enable said interrupt means and said data out gate means during said data output sequence, and to enable said interrupt means and said data in gate means during said data input sequence.

* * * * *

United States Patent [19]

Smith

[11] 3,987,402

[45] Oct. 19, 1976

[54] MULTI-CHANNEL GAIN CONTROLS

[76] Inventor: Russell Peter Smith, 21 Muritai Terrace, Christchurch, New Zealand

[22] Filed: Jan. 15, 1975

[21] Appl. No.: 541,145

[30] Foreign Application Priority Data

Jan. 18, 1974 New Zealand 173129

[52] U.S. Cl. .... 340/1 R; 325/397; 343/7 AG

[51] Int. Cl.² H04B 1/16

[58] Field of Search .... 325/306, 307, 348, 397, 325/398, 401, 407; 340/3 FM, 1 R; 343/5.5 M, 7 AG

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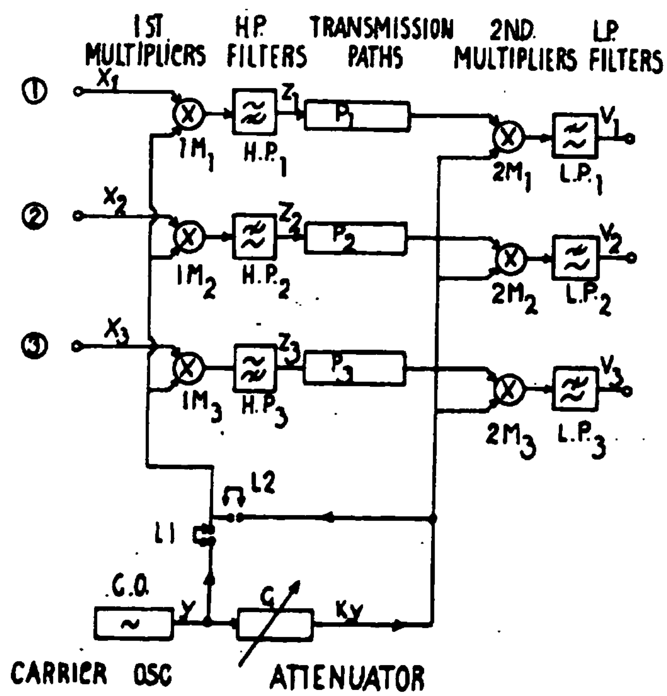
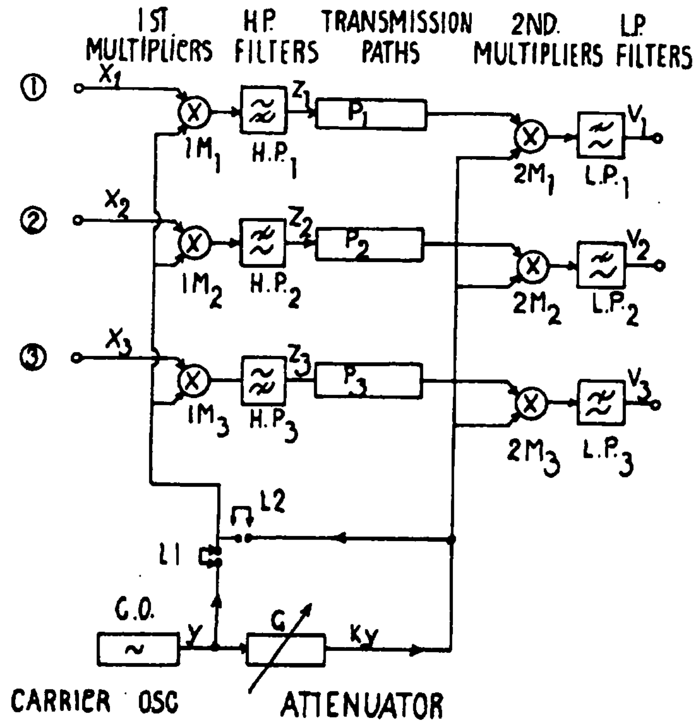
Primary Examiner—Richard A. Farley
Attorney, Agent, or Firm—Holman & Stern

[57] ABSTRACT

A gain control and method therefor which allows the use of a single gang variable attenuator for the synchronous control of the gains of each channel of a multi-channel system of the type employing a separate transmission path for each channel. This is accomplished by inserting multipliers at the output of each transmission path and multiplying the path signals with a carrier signal the level of which is varied with a single gang attenuator. The level of the multiplier output signals vary in accordance with the level of the carrier signal. Where intelligence signals are to be inserted at the input of each transmission path these are multiplied with the carrier signal prior to insertion resulting in the recovery of the intelligence signal at the output of the multipliers located at the output of each path. An increased range of gain control can be achieved by inserting the attenuated carrier signal at the inputs of each transmission path.

13 Claims, 3 Drawing Figures




1ST MULTIPLIERS
H.P. FILTERS
TRANSMISSION PATHS
2ND. MULTIPLIERS
LP FILTERS
1
2
3
X1
X2
X3
1M1
1M2
1M3
H.P.1
H.P.2
H.P.3
Z1
Z2
Z3
P1
P2
P3
2M1
2M2
2M3
L.P.1
L.P.2
L.P.3
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V3
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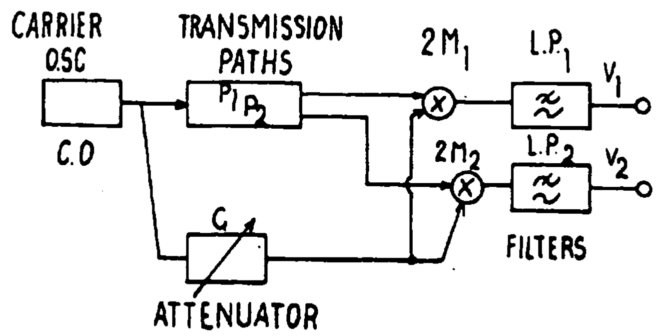
L1
C.O.
Y
G
KY
CARRIER OSC
ATTENUATOR

FIG.1


CARRIER
OSC
C.O
TRANSMISSION
PATHS
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2M1
2M2
L.P.1
L.P.2
V1
V2
FILTERS
G
ATTENUATOR

FIG.2.

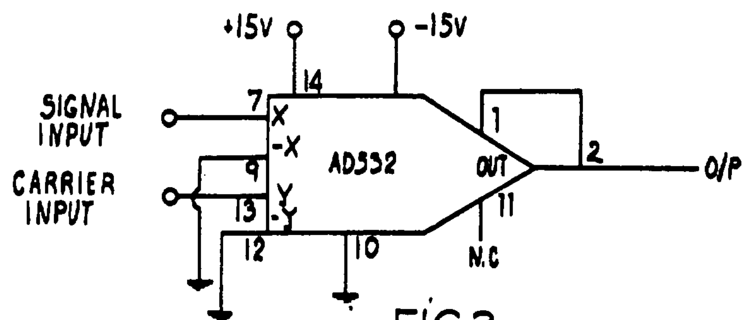
+15V
-15V
SIGNAL
INPUT
CARRIER
INPUT
AD532
OUT
O/P
N.C

FIG.3.

MULTI-CHANNEL GAIN CONTROLS

BACKGROUND OF THE INVENTION

This invention relates to multi-channel gain controls and methods of controlling the gain in a multi-channel system. An example of an application of the invention is as a gain control in radar or sonar systems having more than one receiving channel.

The simplest form of a multi-channel gain control comprises a number of variable attenuators having substantially identical characteristics which are mechanically ganged. To achieve with certainty tracking errors (i.e., gain differences) less than 5% over even a limited dynamic range of gain, it is necessary to use "high precision" controls. Such controls are expensive and bulky. Furthermore, as the maximum attenuation possible is limited by the stray coupling between the attenuator input and output circuits it may be necessary to carry out the attenuation in two or more isolated stages to achieve an adequate dynamic range of gain. When multi-channel attenuation is required with very low tracking errors a wide range of attenuation can only be achieved at great cost and complexity.

An application of a precise multi-channel gain control is in radar or sonar systems having two or more receiving channels to provide source direction information. In such systems it is essential to be able to vary the gains of each channel in precise synchronism, as unequal channel gains will produce directional errors. An example of a sonar system of this type is that described in U.S. Pat. No. 3,366,922 which relates in particular to a binaural sensory aid for blind persons. Two receiving channels are used in these aids and the use of conventional precision ganged attenuators is not compatible with the special packaging requirements.

SUMMARY OF THE INVENTION

It is therefore an object of the present invention to provide a method of controlling the gain of a multi-channel system. It is a further object of the present invention to provide a multi-channel gain control which will go some way towards overcoming the above-mentioned disadvantages. Accordingly in one aspect the invention consists in a method of controlling the gain in a multi-channel system employing a separate transmission path for each channel. A carrier signal is generated and multiplied with the intelligence signal associated with each transmission path. Unwanted low frequencies from each product signal are filtered out and each filtered signal is passed through an associated transmission path. The same carrier signal is also controllably attenuated and the attenuated signal is multiplied with each transmission path output signal. Unwanted high frequencies from these product signals are filtered out. The amplitudes of the intelligence signals resulting after filtering are each determined by the degree of attenuation of the carrier signal.

In a further aspect, the invention consists in a method of synchronously controlling the gain in a multi-channel transmitter-receiver system employing a separate transmission path for each channel. A carrier signal is generated and applied to said transmission paths. This carrier signal is also controllably attenuated and the attenuated signal is multiplied with each transmission path output signal. The unwanted high frequencies from the multiplication product signals are then filtered out. The amplitudes of the signals remaining after filtering are each determined by the degree of attenuation of the carrier signal.

In yet a further aspect the invention consists in a multi-channel gain control for controlling the gain in a system employing a separate transmission path for each channel. A carrier signal oscillator is provided together with a plurality of intelligence signal inputs each associated with a transmission path. First multipliers each feeding a transmission path receive as inputs a respective intelligence signal and the carrier signal from said oscillator, high pass filter means being interposed between the multipliers and the transmission paths to filter out unwanted components of the multiplier outputs. The carrier signal from the oscillator also is taken to a variable attenuator. A plurality of second multipliers each receive as one input the signal from a respective transmission path and as another input the attenuated carrier signal from the attenuator. Low pass filter means follow each second multiplier to filter out unwanted components from the multiplier outputs. The amplitudes of the intelligence signals resulting at the output of the filters are determined by the setting of the attenuator.

In yet a further aspect the invention consists in a gain control for synchronously controlling the gain in a multi-channel transmitter-receiver system employing a separate transmission path for each channel. A carrier signal oscillator is provided and oscillator output means feed the carrier signal to said transmission paths. The output of said oscillator is also applied to a variable attenuator which feeds multipliers corresponding to each transmission path within which the signal from the attenuator is multiplied with the signal received from the respective transmission path. Low-pass filter means filter out unwanted components of the multiplier product signals. The amplitude of the signals at the output of each filter are determined by the setting of the attenuator.

BRIEF DESCRIPTION OF THE DRAWINGS

The preferred forms of the invention will now be described with reference to the accompanying drawings, in which:

FIG. 1 is a block circuit diagram of a multi-channel gain control, including the modification necessary to achieve greater attenuation,

FIG. 2 is a block circuit diagram of a two-channel gain control suitable for use in radar or sonar, and

FIG. 3 is an example of a multiplier suitable for use in the multi-channel gain controls of FIGS. 1 and 2.

DETAILED DESCRIPTION OF THE PREFERRED EMBODIMENTS

The essential feature of the present invention is that the gains in each channel of a multi-channel system may be varied together precisely in synchronism by the use of a single variable attenuator.

Referring to FIG. 1, three channels of a multi-channel system are shown where the respective input intelligence signals are x_1, x_2 and x_3, and the corresponding output signals are v_1, v_2, and v_3. Each channel input signal feeds one of a first set or bank of multipliers $1M_1$, $1M_2$, and $1M_3$. A carrier oscillator C.O. generates a carrier signal y which feeds each of the first multipliers. The output of each multiplier, which may be considered to be the carrier signal y modulated by each intelligence signal x_i (where i is the channel number) is fed to a high pass filter H.P.$_i$ to remove any residual input

signal due to imperfect multiplication. The outputs z_i of the high pass filters feed transmission paths P_i which could vary in nature from a simple connecting lead to an electromagnetic or acoustic transmission medium. The received signal from each transmission path is fed to one of a second bank of multipliers $2M_i$. The carrier signal y, as well as feeding the first multipliers, is taken to a variable attenuator G, and the attenuated output signal K.y feeds each of the second multipliers. The product of the attenuated carrier signal K.y with each received signal is fed to a low pass filter L.P.$_i$, which removes signal components having frequencies greater than the frequencies of the input signals x_i. The resultant output signals v_i are replicas of the input signals x_i diminished by the attenuation factor K. Hence, a variation of the variable attenuator G effects all channels equally.

That the above result is obtained is demonstrated by the following analysis.

Assuming both signal and carrier waveforms are sinusoidal the signal and carrier may be represented respectively as:

$x_i = X_i \sin w_i t$

and

$y = Y \cos w_y t$

where

w_i = the angular frequency of the input signal

w_y = the angular frequency of the carrier signal

X_i = the amplitude of the input signal

Y = the amplitude of the carrier signal

The output of the first multipliers $1M_i$ will be:

$$X_i Y \sin w_i t + S$$

or

$$X_i Y/2 \{\sin(w_i + w_y)t + \sin(w_i - w_y)t\} + S$$

where S represents signals produced by imperfect operation of the multipliers.

Imperfect operation of the multipliers will allow some of the original signals x_i and some of the carrier signal y to appear at the output of each multiplier. High pass filtering will completely eliminate the residual x_i feedthrough if the carrier frequency is chosen appropriately but will allow a residual carrier signal Cy to pass.

The high pass filter output Z_i is then:

$$X_i Y/2 \{\sin(w_i + w_y)t + \sin(w_i - w_y)\} + Cy$$

Any modification by the transmission path can be ignored since this is irrelevant to the operation of the system. The second multiplier output, z_iKy is:

$$X_i K Y^2/4 \{\mathrm{Sin}\,(w_i + 2w_y)t + \sin w_i t + \sin w_i t + (w_i - 2w_y)t\} + CKY^2 (\cos 2w_y t + 1) + S'.$$

Again, due to multiplier limitations, some of z_i and y will also appear at the output. These unwanted outputs are accounted for in the term S'. However, providing the carrier signal has been chosen to be greater than twice the frequency of x_i the only terms which appear at the output of the LP filters are:

$$v_i = \frac{X_i K Y^2}{2} \sin W_i t + \frac{CKY^2}{2}$$

The second term is a small spurious DC term which may be minimised by optimum adjustment of the first multiplier and in many applications is irrelevant.

The useful output term is:

$v_i = ½ X_i K Y^2 \sin w_i t$

$= K/2\ Y^2 x_i$

$v_1 = K/2\ Y^2 x_1$

$v_2 = K/2\ Y^2 x_2$

$v_3 = K/2\ Y^2 x_3$

That is, the amplitudes of signal outputs of all channels vary directly with K.

With the above method of gain control tracking between channels to within 0.1 dB over a dynamic range of 60 dB can be achieved without difficulty.

Various known multipliers, filters, oscillators and variable attenuators may be used in a realisation of the block circuit shown in FIG. 1. However, as an example, a typical multiplier circuit is shown in FIG. 3. The AD 532 multiplier used is an integrated circuit of the 14 pin package type produced by Analogue Devices Incorporated. The variable attenuator may be a precision step switched type, or a simple variable resistance type, depending on the application. Similarly, the characteristics of the high and low pass filters will vary according to the application.

In a second form of the invention a considerably greater range of attenuation is provided by applying the attenuated carrier signal Ky to each multiplier of the first bank of multipliers. This modification is illustrated diagramatically in FIG. 1 where link L1 would be removed and link L2 inserted. In this case, the carrier signal feeding the first multipliers also reduces as K reduces. The output signals of this form of the invention may be represented as:

$$V_i = K^2/2\ Y^2 x_i$$

Thus, the outputs now all vary directly with K^2. In this second form of the invention the dynamic range over which the gain may be varied is greatly increased due to the effective provision of two separate stages of attenuation, each having the same range as the single stage of the first form of the invention. As previously mentioned, the range of attenuation is limited by the stray coupling between the input and the output of the attenuator or direct coupling of the signal source into the output circuit. This two-stage attenuation approach as well as providing a theoretically greater range of attenuation, reduces coupling between the input and output by use of two isolated stages of attenuation. In terms of decibels the second form of the invention provides twice the attenuation of the first form.

A further form of the invention will now be described with reference to a particular application.

In certain types of sonar and radar systems a transmitted signal is frequently used in conjunction with two or more receivers to derive directional information of a source of transmitted signal reflection. The direction information may be given by differences in the relative levels of the received signals. For example in the binaural sensory aid described in U.S. Pat. Nos. 3,366,922 the transmitted signal is frequency swept and accordingly the received echo along two different paths will show a difference in frequency with respect to the transmitted signal at any one instant. The received signals are therefore each multiplied with the transmitted carrier to produce audible signals having frequencies equal to the difference between the received frequency of each echo and the transmitted signal frequency. The beat frequencies are indicative of the distance between each receiver and the source of reflection.

The transmitting and receiving transducers have directional characteristics such that directions information on the source of reflecti n is supplied by the differing relative levels f the two received signals. In such a system it is essential to be able to contr l the gains of the receiver channels in precise synchronism since differing channel gains will lead to errors in the derived direction information.

Since the sensory aid uses two acoustic/electric receiving transducers mounted with a single electric/acoustic transmitting transducer the system is a two channel one, each transmission path comprising the air path from the transmitting device to a reflecting object and the return air path to a respective receiving device. The useful information derived from each receiving device is the time taken for the acoustic signal to traverse each transmission path and the angle each path makes with the receiving transducers. Because the "intelligence" information in this system is inherently produced by the transmission paths the carrier signal feeds both transmission paths without prior multiplication with external intelligence signals. As will be appreciated from FIGS. 1 and 2 and associated description of U.S. Pat. No. 3,172,075 which describes the electronics used in the aid of U.S. Pat. No. 3,366,922, the transmission path reflection time information is made available by frequency sweeping the transmitted carrier signal and multiplying this frequency swept signal with the received and delayed reflections of the transmitted signal. Accordingly the gain control described with reference to FIG. 1 of the present specification takes the configuration shown in FIG. 2 when used to control the amplitudes of the output signals of the two channel blind aid system. This can be compared with FIG. 1 of U.S. Pat. No. 3,172,075 and it will be seen that the modification permitting synchronous gain control of both receiving units is the provision of attenuator G in the carrier signal feeds to the multipliers (labelled frequency changes 19 in the U.S. Patent). The level of the audible signal from both multipliers thus varies in accordance with the attenuation factor K of attenuator G.

The input signal to the transmission paths would be obtained from an electric/acoustic transducer and the output signals of said paths would be applied to electric/acoustic transducers.

Again, in correspondence with the first form of the invention, to enable an increased range of gain control the signal feeding the transmission paths can be that at the output of the attenuator.

I claim:

1. A method of controlling the gain in a multi-channel system employing a separate transmission path for each channel comprising:
  generating a carrier signal,
  multiplying each of a plurality of intelligence signals each associated with a transmission path with said carrier signal to produce first product signals,
  filtering unwanted low frequencies from each first product signal, passing each filtered signal through an associated transmission path,
  controllably attenuating said carrier signal,
  multiplying the signal from each transmission path with the attenuated carrier signal to produce second product signals,
  filtering unwanted high frequencies from each second product signal, the amplitudes of the intelligence signals resulting after filtering each being determined by the degree of attenuation of the carrier signal.

2. A method according to claim 1, wherein the carrier signal multiplied with the intelligence signal is said attenuated carrier signal.

3. A method of synchronously controlling the gain in a multi-channel transmitter-receiver system employing a separate transmission path for each channel comprising:
  generating a carrier signal,
  controllably attenuating said carrier signal,
  applying the attenuated carrier signal to the inputs of said transmission paths,
  multiplying the signal from the output of each transmission path with the attenuated carrier signal, and filtering out unwanted low frequencies from the multiplication product signals,
  the amplitudes of the signals remaining after filtering each being determined by the degree of attenuation of the carrier signal.

4. A multi-channel gain control for controlling the gain in a system employing a separate transmission path for each channel comprising:
  a carrier signal oscillator,
  a plurality of intelligence signal inputs each associated with a transmission path,
  a plurality of first multipliers each feeding a transmission path and each receiving as inputs an intelligence signal and the carrier signal from said oscillator,
  high pass filter means interposed between said multipliers and said transmission paths to filter out unwanted components of the multiplier outputs,
  a variable attenuator to which the output of said oscillator is applied,
  a plurality of second multipliers each receiving as one input the signal from a transmission path and as another input the attenuated carrier signal from said attenuator, and low pass filter means following each second multiplier to filter out unwanted components from the multiplier output,
  the amplitudes of the intelligence signals resulting at the output of the filters each being determined by the setting of said attenuator.

5. A multi-channel gain control according to claim 4 wherein the carrier signal input to the first multipliers is taken from the output of said attenuator.

6. A multi-channel gain control according to claim 4 wherein each transmission path is an electrical conductor.

7. A multi-channel gain control according to claim 5 wherein each transmission path is an electrical conductor.

8. A multi-channel gain control according to claim 4 wherein the carrier signal oscillator frequency is greater than twice the intelligence signal frequency.

9. A gain control for synchronously controlling the gain in a multi-channel transmitter-receiver system employing a separate transmission path for each channel comprising:
  a carrier signal oscillator,
  a variable attenuator to which the output of said oscillator is applied,
  oscillator output means feeding the attenuated carrier signal to the inputs of said transmission paths,
  a plurality of multipliers each corresponding to a transmission path within which each signal received from the output of a respective transmission

path is multiplied with the attenuated carrier signal,

and low pass filter means to filter out unwanted components of the product signals, the amplitudes of the signals at the output of the filters each being determined by the setting of said attenuator.

10. A gain control according to claim 9 wherein the number of channels controlled is two.

11. A gain control according to claim 10 wherein the carrier signal oscillator frequency is in the ultrasonic range.

12. A gain control according to claim 11 wherein the transmissi n paths are acoustic media, the oscillator output means is an electric/acoustic transducer, and an electric/acoustic transducer is associated with and feeds each multiplier.

13. A gain control according to claim 12 wherein the transmission path transmitter and receiver terminals are located together.

* * * * *

United States Patent [19]

Smith

[11] **4,067,013**

[45] **Jan. 3, 1978**

[54] **AUTOMATIC THRESHOLDING AND REFERENCE CIRCUIT**

[75] Inventor: **Richard M. Smith, Ridgecrest, Calif.**

[73] Assignee: **The United States of America as represented by the Navy, Washington, D.C.**

[21] Appl. No.: **741,261**

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[51] Int. Cl.[2] G01S 9/42; G01S 7/34; H04B 1/12

[52] U.S. Cl. .... 343/7.7; 343/17.1 R; 325/474

[58] Field of Search .... 343/7 A, 7.7, 17.1 R; 340/15.5 A, 15.5 AP, 15.5 AC, 15.5 BH; 325/474, 477

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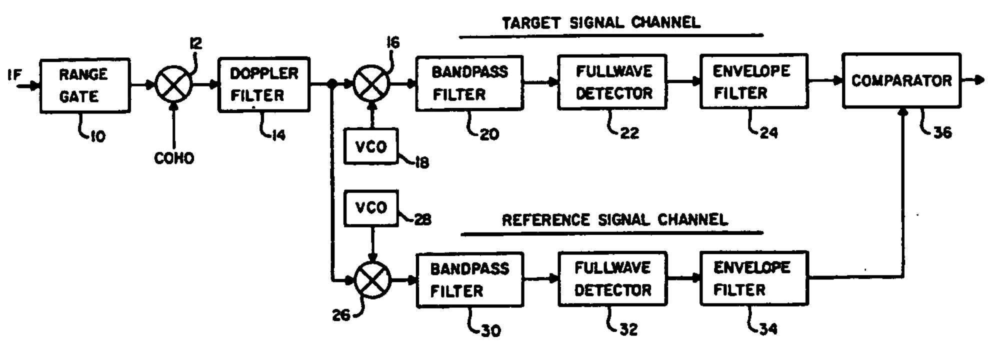
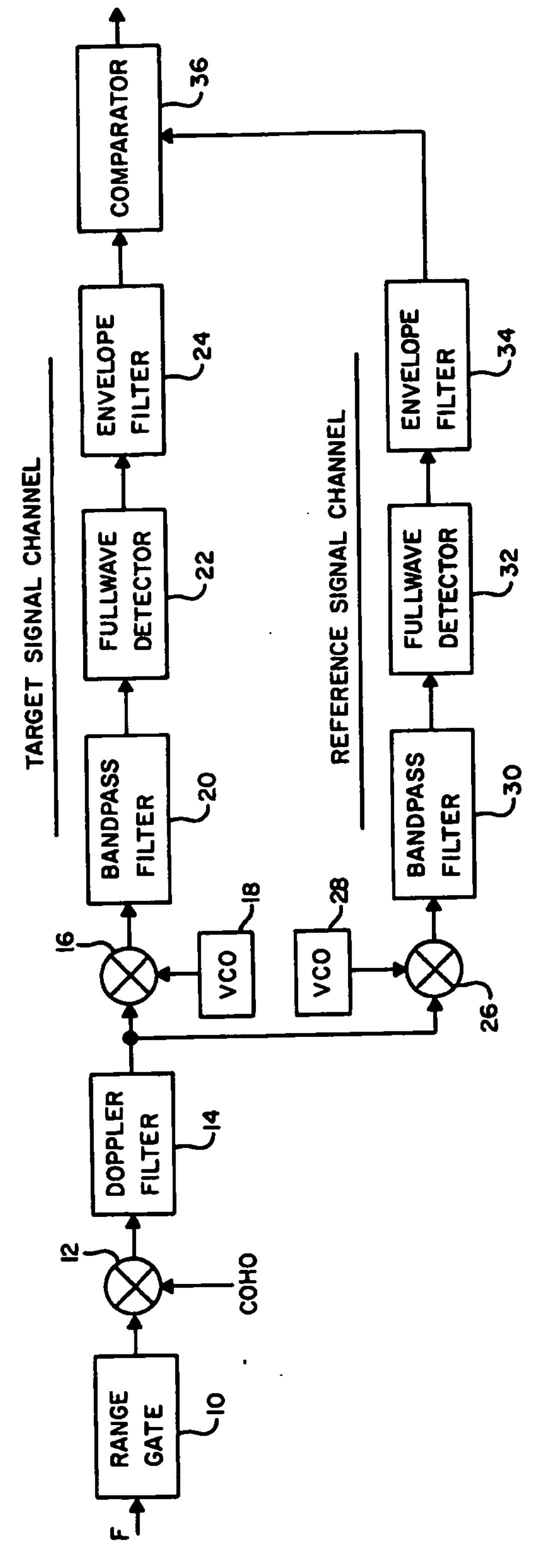
Primary Examiner—Maynard R. Wilbur
Assistant Examiner—Lawrence Goodwin
Attorney, Agent, or Firm—R. S. Sciascia; Roy Miller

[57] **ABSTRACT**

An automatic thresholding reference and detection circuit which generates a signal in time which is a replica of the radar return signal's doppler spectrum. The radar signal return is processed by a target signal channel having a narrow bandwidth filter which produces a fine resolution replica of the signal spectrum, and by a reference signal channel having a wider bandwidth filter, which produces a smeared (or average) replica of the signal spectrum. The output of the reference signal channel is a threshold signal which, when compared to the output of the target signal channel, allows only the target signals to be passed for further processing and rejects clutter and thermal noise.

9 Claims, 8 Drawing Figures




IF
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10
12
COHO
DOPPLER FILTER
14
16
VCO
18
BANDPASS FILTER
20
FULLWAVE DETECTOR
22
ENVELOPE FILTER
24
TARGET SIGNAL CHANNEL
COMPARATOR
36
VCO
28
26
BANDPASS FILTER
30
FULLWAVE DETECTOR
32
ENVELOPE FILTER
34
REFERENCE SIGNAL CHANNEL

Fig. 1

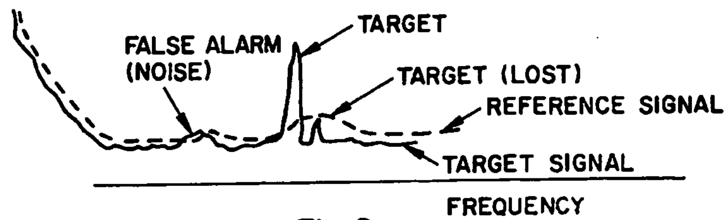

Fig. 2 a

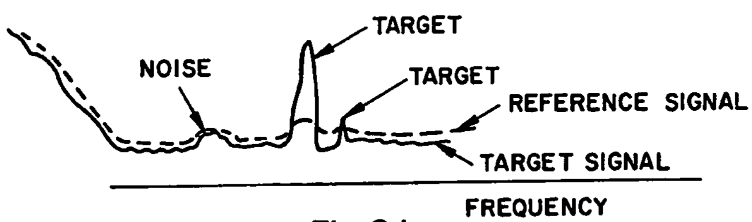

Fig. 2 b

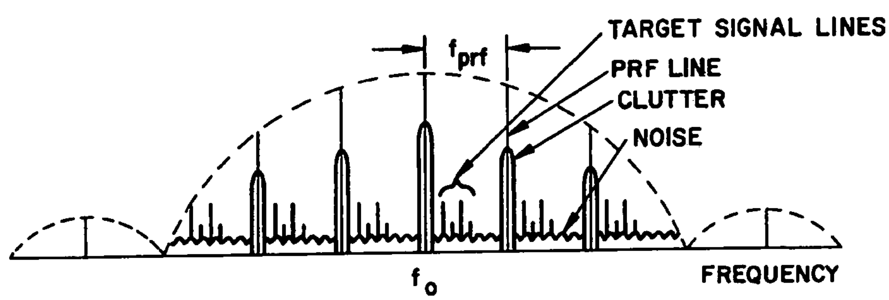

Fig. 3 a

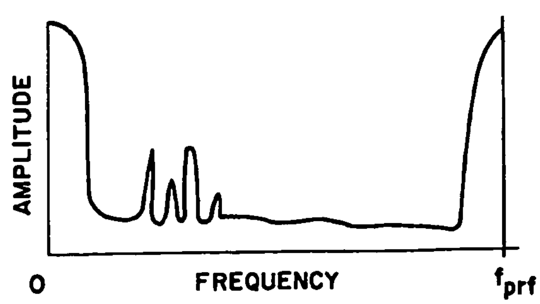

Fig. 3 b

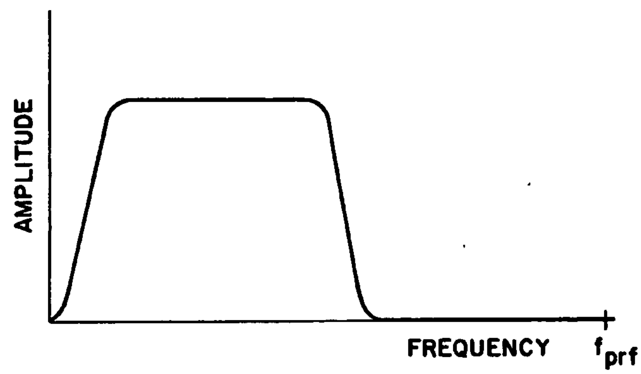

Fig. 4 a

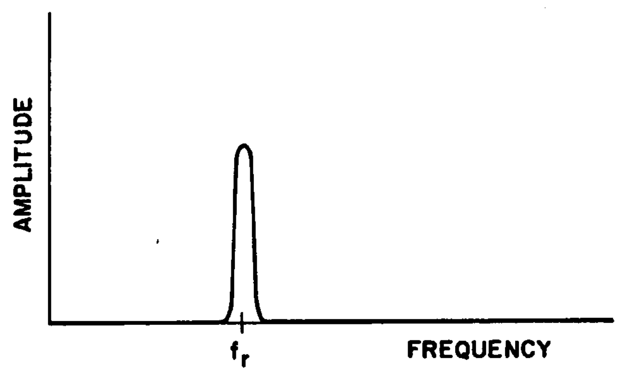

Fig. 4 b

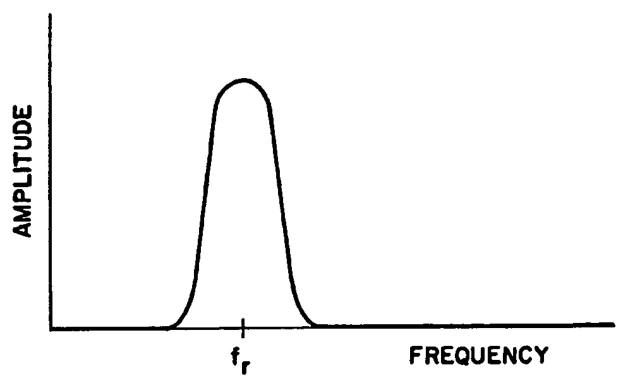

Fig. 4 c

AUTOMATIC THRESHOLDING AND REFERENCE CIRCUIT

BACKGROUND OF THE INVENTION

The present invention relates to Doppler radar systems, and more particularly to an automatic thresholding circuit for Doppler radar systems.

The type of target to be tracked by a system which incorporates this invention has a "line-like" Doppler frequency spectrum. For example, for aircraft targets, the energy of the target returns is divided between the skin line and a series of upper and lower sideband lines generated by engine modulations. These upper and lower sideband lines may have more amplitude, for short periods, than the skin line.

One way of suppressing false alarms is to employ some type of automatic thresholding. Automatic thresholding in pulse doppler radars is done in several ways. One of the most common methods is to provide a bank of doppler filters to cover the frequencies between the pulse repetition frequency (PRF) lines. Detectors follow each doppler filter. The signal levels are compared against a threshold. If the signal level is greater than the threshold then the signal is detected. This threshold level can be generated by averaging the total power in the pass band between the PRF lines or by averaging noise power for a subgroup of the doppler filters.

Another common method uses returns from neighboring range bins about the range bin to be examined. These returns are processed individually as in the previous method and then added to form a threshold.

These methods are practical where component and processing complexity can be tolerated. Depending upon the actual implementation, these thresholding methods may not be very sensitive to variations in noise power in the intra PRF line frequency region caused by clutter and transmitter noise.

SUMMARY OF THE INVENTION

Accordingly, the present invention provides two parallel processing channels — a target signal channel and a reference signal channel to provide automatic thresholding and detection. After range gating, the received return signal spectrum is shifted to a low frequency in a mixer. The doppler frequency region of interest within the return spectrum between two PRF lines is passed through a doppler band pass filter. The doppler frequencies are then converted to time signal replicas of the spectrum by two parallel signal channels. The parallel signal channels contain mixers, narrow band pass filters, fullwave envelope detectors and an envelope filter.

The bandwidth of the narrowband filter for the reference signal channel is wider than that for the target signal channel to provide an average reference value. The VCO for the reference signal channel is offset from that of the target signal channel to compensate for the different rise times of the two different narrow bandwidth filters. The VCO offset allows for centering the fine resolution time replica of the signal spectrum with that of the coarser replica used for reference.

The output of the two channels is compared and the target signals are passed for further processing.

BRIEF DESCRIPTION OF THE DRAWING

FIG. 1 is a block diagram of the present invention;

FIG. 2 is a diagram illustrating the effect upon the reference signal (a) without and (b) with an offset VCO in the reference signal channel;

FIG. 3 is a diagram illustrating (a) the received return frequency spectrum and (b) the processed frequency spectrum; and

FIG. 4 is a diagram of the respective bandpass characteristics of (a) the doppler filter, (b) the target signal filter, and (c) the reference signal filter.

DESCRIPTION OF THE PREFERRED EMBODIMENT

Referring now to FIG. 1, an IF output signal from a pulse doppler radar is fed to a range gate 10. From the range gate 10 the signal is down-translated in frequency by mixing the output of the range gate with the output of a coherent oscillator (COHO) in a mixer 12. The output of the mixer 12 is then inputted to doppler filter 14 which passes only the frequency region of interest between two PRF lines. The signal from the doppler filter 14 is then fed as one input to another pair of mixers 16, 26. A pair of voltage controlled oscillators (VCO) 18, 28 drive mixers 16, 26 respectively, with a linearly swept frequency in time.

The output from the doppler filter 14 interacts with the outputs of the VCO's 18 and 28 in the respective mixers 16, 26 to shift the doppler signal spectrum in time. The output of the mixers 16, 26 is then coupled through narrow bandpass filters 20, 30 with the respective bandwidths of the filters centered at the same frequency. The bandwidth of the filter 30 in the reference signal channel is wider than that of filter 20 in the target signal channel.

The respective signals, target and reference, are then processed by full-wave detectors 22, 32 and envelope filters 24, 34. The outputs of the two envelope filters 24, 34 provide the replica of the envelope of the signal spectrum as a function of time. These two signals are then coupled to a comparator 36 where the reference signal acts as a threshold signal so that only target signals are passed for further processing.

Bandpass filters 30 and 20 have different rise times and compensation for this is provided by offsetting VCO 28.

The effect of the offset VCO 28 is illustrated in FIG. 2. FIG. 2*a* illustrates that without the offset VCO, noise can be passed as a target, creating false alarms. Also, a low amplitude target signal next to a high amplitude target signal could be lost. Raising the reference level could reduce the false alarms, but sensitivity would be sacrificed. However, by offsetting VCO 28 the results of FIG. 2*b* are obtained, i.e., the noise is suppressed and the low amplitude target is detected without reducing sensitivity.

This method of providing automatic thresholding discriminates against non line-like spectrums. The narrow reference filter 30 provides an average reference signal over only a portion of the intra PRF line frequencies. This provides a better reference if the clutter and noise power changes significantly across the frequency band examined and allows a more sensitive signal to be detected.

The radar used in the present invention is coherent and the narrow filtering of the return spectrum can therefore provide processing gain. For example, if the PRF lines are separated by 50kHz and the effects of clutter are excluded, the signal-to-noise ratio, S/N, for the signal and n ise energy between the PRF lines is the

same as for the total signal. If the narrow bandpass filter 20 is placed over a target signal, all of the signal is passed and only a small portion of the noise. Increasing the S/N ratio in this way provides a processing gain increase on the order of approximately 18.8db in the present design.

Referring now to FIG. 3*a*, a typical signal return spectrum is illustrated centered about the transmitter frequency, f_0. The return signal consists of a series of PRF lines with clutter roughly centered about them, and target signal lines between the PRF lines with a thermal noise background. If the range gate 10 moves off a target, the only change in the shape of the signal spectrum between the PRF lines is in amplitude. The range gated signals are mixed with a COHO signal to down-translate the frequency spectrum from the IF frequency to d.c. as shown in FIG. 3*b*.

The doppler filter 14 passes the signal frequencies between d.c. and the first PRF line and rejects the frequencies of the stationary clutter signals centered about the d.c. and PRF lines. Then the likely doppler target signal frequencies between d.c. and the first PRF line are searched for signal energy. This is accomplished by sweeping the signal spectrum past the narrow bandpass filter 20 fixed in frequency.

FIG. 4 shows the respective bandpass characteristics of the doppler filter 14 (FIG. 4*a*), the target signal narrow bandpass filter 20 (FIG. 4*b*), and the reference signal narrow bandpass filter (FIG. 4*c*) with typical bandwidths of 20kHz, 660Hz and 3KHz, respectively. The target signal filter 20 and the reference signal filter 30 have their respective bandwidths centered about frequency f_r.

The output of the doppler filter 14 is multiplied by the VCOs signals which are changing linearly in frequency. Since multiplication of time signals is equivalent to convolution in frequency, the signal spectrum slides through the narrow bandpass filters 20, 30 centered at f_r. The VCO sweep rate determines the bandwidth of the target signal filter 20 required to provide enough dwell time to accomodate the coherent integration of the target signal lines. The VCO signals are designed so that the spectrum shifted into the narrow bandpass filters 20, 30 stops just short of the shifted d.c. signal to provide further clutter attenuation.

The reference signal is derived from the doppler filter 14 to provide for the adaptive threshold to the comparator 36. The reference signal is centered with respect to the target signal to compensate for the difference in rise times of the reference signal filter 30 and target signal filter 20.

The comparator 36 controls a transmission gate which passes any signal which is of greater amplitude than the reference signal.

The present invention overcomes the slow time constants and resulting offset time response of narrow bandpass filters by using an offset VCO 30 in the reference signal channel, tracks changing noise levels within the doppler frequency spectrum of interest between two PRF lines, discriminates against detecting non line-like spectrums, and allows for simple mechanization.

What is claimed is:

1. An automatic, adaptive-threshold circuit for processing signals received from a pulse Doppler radar comprising:

reception means for receiving a radar return signal from said pulse Doppler radar;

a target channel connected to the utput of said reception means, tuned to a first narrow bandwidth, and utputting a first time replica spectrum f said received output from said recepti n means;

a reference channel connected t the output of said reception means and in parallel with said target channel, said reference channel tuned to a second narrow bandwidth wider than said first narrow bandwidth and outputting a second time replica spectrum of said received output from said reception means, said second time replica spectrum being coarser than said first time replica spectrum; and

a comparator having a plurality of inputs connected to and receiving the outputs of said target channel and said reference channel, whereby said received radar return signal is processed in parallel fashion by said target and said reference channels, said reference channel output being the automatic, adaptive threshold against which said target channel output is compared in said comparator.

2. An automatic, adaptive threshold circuit as recited in claim 1 wherein said comparator controls a transmission gate which passes only that portion of the output of said target channel which is of greater amplitude than the output of said reference channel.

3. An automatic, adaptive threshold circuit as recited in claim 1 wherein said reception means includes:

gating means connected to receive said radar return signal for range-gating said radar return signal;

lowering means connected to the output of said gating means for down-translating in frequency the output of said gating means; and

filter means connected to the output of said lowering means for selecting only the Doppler frequencies to be processed.

4. An automatic, adaptive threshold circuit as recited in claim 3 wherein said lowering means includes:

a coherent oscillator having an output oscillation signal; and

a first mixer having inputs connected to said gating means and said coherent oscillator and outputting a signal to said filter means.

5. An automatic, adaptive threshold circuit as recited in claim 1 wherein said target channel includes:

a second mixer having a plurality of inputs, one of which is connected to the output of the aforesaid filter means;

a first voltage-controlled oscillator having a frequency output swept linearly in time which is connected to another input of said second mixer;

a first narrow-bandpass filter connected to the output of said second mixer for establishing said first narrow bandwidth;

a first full-wave detector connected to the output of said first narrow-bandpass filter; and

a first envelope detector connected to the output of said first full-wave detector, said first envelope detector outputting a signal to one input of the aforesaid comparator.

6. An automatic, adaptive threshold circuit as recited in claim 5 wherein said first narrow bandpass filter is fixed in frequency.

7. An automatic, adaptive threshold circuit as recited in claim 1 wherein said reference channel includes:

a third mixer having a plurality of inputs, one of which is connected to the output of the aforesaid filter means;

a second voltage-controlled oscillator having a frequency output swept linearly in time which is connected to an ther input f said third mixer;

a second narrow-bandpass filter connected to the output f said third mixer for establishing said second narrow bandwidth, said second narrow bandwidth being wider than said first narrow bandwidth;

a second full-wave detector connected to the output of said second narrow-bandpass filter; and

a second envelope detect r connected t the utput f said second full-wave detector, said second envel pe detect r utputting a signal to an ther input of the aforesaid comparator.

8. An automatic adaptive thresh ld circuit as recited in claim 7 wherein said second narrow-bandpass filter is fixed in frequency.

9. An automatic, adaptive threshold circuit as recited in claim 7 wherein said second voltage-controlled oscillator is offset in time from said first voltage-controlled oscillator.

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